

ACTIVE CONTROL OF NOISE
RADIATED FROM PERSONAL COMPUTERS

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(ABSTRACT)

One of the most critical issues in computer or semi-conductor design is heat dissipation. It is also the primary source of failure of electronic equipment. Thus, a major effort in the computers design is to ensure proper cooling of the processor units. Due to the heat increase with modern chips, cooling fans have become larger in size and have faster speed. Very often, two to three fans are incorporated in PC units for cooling purposes. As an indirect consequence of increased heat cooling requirements, computers have thus become noisier and it is expected that the sound quality of those machines will become poorer. Mechanical data storage devices such as hard disk drives also contribute to the annoying noise radiated by personal computers. Thus, a need for control of computer noise has arisen.

An important challenge in personal computers is the size constraint, which makes the implementation of passive noise control techniques very difficult. Consequently, a study of the potential of active noise control (ANC) to develop a cost effective, compact solution to the noise problems discussed above was proposed. This technique seems particularly well suited to computer noise since most of the hardware required for its implementation is present in contemporary personal computers. Moreover, commercial applications of ANC will increase rapidly in the near future as the price of digital signal processors (DSP), analog to digital (A/D), and digital to analog (D/A) converters continues to fall.

The first goal of this work was to identify and characterize the major stationary noise sources in computers. The second goal was to globally control the computer noise using hybrid passive / active technology while the last goal was to implement a purely active

local noise control strategy. An investigation of the major noise radiation paths helped to define a strategy for the combination of passive and active techniques to control the PC noise. Passive redesign of the PC casing aimed at reducing both structure-borne as well as airborne noise radiation paths. The structure-borne path was altered through the decoupling of the vibrating sources from the chassis. The implementation of folded lined ducts covering the PC air inlet and outlet lead to the control of most of the PC airborne noise above 1000Hz. Active control of fan noise in short ducts was then investigated using a single input / single output feedforward adaptive digital controller based on the Filtered-X LMS algorithm. Such control strategy lead to up to 10dB of broadband noise attenuation at the outlet of the ducts. However, the active / passive sound absorber in ducts did not lead to a significant global reduction of the PC noise. It was proposed but not confirmed that such limited performance is due to important airborne noise radiation through the casing and leaks. Finally, local control of PC noise at the user head location in a simulated semi-reverberant environment was investigated using a multiple input / multiple output feedforward adaptive digital controller based on the Filtered-X LMS algorithm. A large zone of quiet surrounding the head was created at low frequencies (6 to 18dB of pure tone reduction at 250Hz in a 12 by 12 by 18 inches zone surrounding the head) when combining two control sources (small amplified bookshelf speakers) and two error sensors (B&K microphones). Nevertheless, the zone of quiet would reduce in size with increasing frequency, eventually leading to zones of constructive interference for frequencies above 1000Hz.

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à mes parents,

Contents

ACKNOWLEDGMENT.....	IV
CONTENTS	VI
LIST OF FIGURES.....	IX
LIST OF TABLES	XV
CHAPTER 1. INTRODUCTION.....	1
1.1. Motivation.....	1
1.2. Active Noise Control.....	2
1.3. Objectives and Organization	8
CHAPTER 2. MEASUREMENTS AND CONTROL TECHNIQUES REVIEW	10
2.1. Signal Processing Tools	10
2.1.1. Power Spectrum Estimates	10
2.1.2. Linear SISO Systems Study and Application to ANC.....	16
2.1.3. Linear MISO System Study and Application to ANC.....	20
2.1.4. Delay Estimation.....	26
2.2. Acoustical Measurements of Noise Sources.....	33
2.2.1. Acoustic Power Level	34
2.2.2. Intensity Survey	39
2.3. Active Noise Control.....	45
2.3.1. Adaptive Systems.....	49
2.3.2. Practical Implementation of Adaptive Algorithms	58
CHAPTER 3. NOISE SOURCES IDENTIFICATION	74
3.1. Dell Precision Workstation 410 Computer.....	75
3.1.1. Geometry.....	75
3.1.2. Hardware Components.....	76
3.1.3. Identified Noise Sources	77
3.2. Sound Power Level Measurements	77

3.2.1. Experimental Setup	78
3.2.2. Results	84
3.3. Directivity Survey	103
3.3.1. Experimental Setup and Data Processing	104
3.3.2. Results	105
3.4. Overall Noise Radiation	110
3.4.1. Intensity Survey on the Casing	110
3.4.2. Ranking of the Noise Sources	118
CHAPTER 4. PASSIVE CONTROL OF PC NOISE	126
4.1. Structural Decoupling of Vibrating Noise Sources	127
4.1.1. Processor Fan Noise	127
4.1.2. Hard Drive Noise	129
4.2. Fan Ducting	131
4.2.1. Duct Acoustics	131
4.2.2. First Duct Prototype	145
4.2.3. Second Duct Prototype	153
CHAPTER 5. GLOBAL ACTIVE NOISE CONTROL	172
5.1. Review of the Filtered-X LMS Feedforward Control System	173
5.2. Implementation of ANC in Ducts	175
5.2.1. Ducts: a Privileged Global ANC Environment	175
5.2.2. ANC Architecture	177
5.2.3. Key Parameters for the Controller Performance	179
5.3. Practical Design	182
5.3.1. First Duct Prototype	182
5.3.2. Second Duct Prototype	214
CHAPTER 6. LOCAL ACTIVE NOISE CONTROL	232
6.1. Control Setup Presentation	233
6.1.1. Single or Multiple Channel Controller?	233
6.1.2. Implementation in a Personal Computer	234
6.1.3. Operating Environment	235
6.2. SISO Feedforward Control Results	236
6.2.1. Reference Signal Selection	236
6.2.2. Control Results	243

6.3. MIMO Feedforward Control Results	256
6.3.1. Reference Signal Selection	256
6.3.2. Control Results.....	266
 CHAPTER 7. CONCLUSIONS AND FUTURE WORK.....	 296
7.1. Conclusions.....	296
7.2. Future Work.....	300
 REFERENCES.....	 302
 VITAE	 304

List of Figures

Figure 1.1. Illustration page of Paul Lueg's 1934 patent [4]	4
Figure 1.2. Illustration of the work of Olson and May [5].	4
Figure 1.3. Generic control architecture.....	7
Figure 2.1 Multiple input / single output model for arbitrary inputs	20
Figure 2.2 Multiple input / single output model for ordered conditioned inputs	23
Figure 2.3 Sound power level measurement in one of the VAL reverberation chambers	37
Figure 2.4 Sound power level measurement in one of the VAL anechoic chambers	38
Figure 2.5 Sound power level measurement in one of the VAL anechoic chambers	45
Figure 2.6 Generic control architecture.....	47
Figure 2.7 Adaptive feedforward controller architecture	48
Figure 2.8 Generic block diagram of a digital filter.....	50
Figure 2.9 Block diagram of an L coefficients FIR digital filter	51
Figure 2.10 Block Diagram of an adaptive filter.....	52
Figure 2.11 MSE surface. Simple case where $L=2$	55
Figure 2.12 System identification block diagram	58
Figure 2.13 Single channel broadband feedforward ANC in duct	59
Figure 2.14 Complete block diagram of the single channel feedforward ANC in duct.....	60
Figure 2.15 Secondary path modeling in single channel feedforward ANC.....	61
Figure 2.16 Block diagram of a single channel feedforward controller based on FXLMS	63
Figure 2.17 Secondary path identification block diagram	64
Figure 2.18 Block diagram of ANC system with noise on the input signals	65
Figure 2.19 Block diagram of control system with feedback effect	66
Figure 2.20 Simultaneous identification of the secondary and feedback paths	67
Figure 2.21 SISO feedforward ANC in duct. FXLMS with feedback removal filter	68
Figure 2.22 MIMO Filtered-X LMS control system.....	69
Figure 3.1 Overall view of the Dell Precision Workstation 410 casing.....	75
Figure 3.2 Hardware configuration in the Workstation 410	76
Figure 3.3 Sound power level measurement in one of the VAL anechoic chambers	78
Figure 3.4 Sound power level measurement in a VAL reverberant chamber	79

Figure 3.5 A-weighting filter gain versus frequency curve.....	83
Figure 3.6 Power supply block with integrated fan.....	84
Figure 3.7 Power supply block sound power level. Repeatability issue.	86
Figure 3.8 Casing effect on power supply block noise (anechoic chamber measurements).....	87
Figure 3.9 Casing effect on power supply block noise (reverb. chamber measurements).....	88
Figure 3.10 Chassis mounted processor fan.....	89
Figure 3.11 Processor fan sound power level. Repeatability issue.	91
Figure 3.12 Casing effect on the processor fan radiation.....	92
Figure 3.13 Baffled processor fan testing in the anechoic chamber.	93
Figure 3.14 Processor fan sound power level. Bipolar characteristics.....	94
Figure 3.15 Processor fan blades protection.	95
Figure 3.16 Processor fan sound power level. Protection screen effect.....	95
Figure 3.17 Processor fan sound power level. Mounting effect.....	97
Figure 3.18 Internal hard drive.....	99
Figure 3.19 Hard drive sound power level. Repeatability issue.....	101
Figure 3.20 Casing effect on the Hard drive radiation.....	102
Figure 3.21 Hard drive sound power level. Mounting effect.	103
Figure 3.22 Directivity measurements in a VAL anechoic chamber.	104
Figure 3.23 Power supply block directivity	105
Figure 3.24 Un-baffled processor fan directivity	106
Figure 3.25 PC casing vertical directivity: front panel.	108
Figure 3.26 PC casing vertical directivity: rear panel.....	108
Figure 3.27 PC casing horizontal directivity: front panel.	109
Figure 3.28 PC casing horizontal directivity: rear panel.....	109
Figure 3.29 Intensity survey on the PC in a VAL anechoic chamber.....	112
Figure 3.30 Power spectrum integrated over the front, rear and side panels.	113
Figure 3.31 Intensity scan over the PC casing. Frequency range: 100-6500Hz.	113
Figure 3.32 Intensity map over the PC casing. Third octave band: 250Hz.....	115
Figure 3.33 Intensity map over the PC casing. Third octave band: 500Hz.....	116
Figure 3.34 Intensity map over the PC casing. Third octave band: 1260Hz.....	117
Figure 3.35 Intensity map over the PC casing. Third octave band: 2520Hz.....	117

Figure 3.36 PC overall sound power level. Repeatability issue.....	119
Figure 3.37 PC overall sound power level. Broad frequency range.....	120
Figure 3.38 PC overall sound power level. Broad frequency range.....	121
Figure 3.39 User environment simulation in a VAL anechoic chamber.....	123
Figure 3.40 SPL at head location. Simulated user environment.	125
Figure 4.1 Chassis mounted processor fan.....	128
Figure 4.2 Processor fan structural decoupling effect on PC sound power level.....	128
Figure 4.3 Hard drive structural decoupling effect on PC sound power level.	129
Figure 4.4 Hard drive mounting in standard bracket.....	130
Figure 4.5 Waveguide modeling.	132
Figure 4.6 Different designs of dissipative silencers.	138
Figure 4.7 Parallel-baffle type silencer	141
Figure 4.8 Illustration of multiple reflections phenomenon in a folded duct.....	144
Figure 4.9 Compromises between thermal and acoustic performance in lined ducts.	146
Figure 4.10 Picture of the first outlet duct prototype.	147
Figure 4.11 Half front view the first outlet duct prototype.	148
Figure 4.12 Section cut of the first outlet duct prototype.....	149
Figure 4.13 Pseudo insertion loss measurement setup.....	150
Figure 4.14 Pseudo insertion loss results.	152
Figure 4.15 Illustration of the calculation of liner percentage in a duct.	155
Figure 4.16 Picture comparing the old and new speakers used in the ducts.	155
Figure 4.17 Overall view of the PC with the second duct prototype.	157
Figure 4.18 View of the inlet duct with the structurally decoupled processor fan.....	157
Figure 4.19 Half front view of the second duct prototype (outlet).....	158
Figure 4.20 Section cut of the second duct prototype (outlet).	159
Figure 4.21 Front view of the second duct prototype (inlet).....	160
Figure 4.22 Section cut of the second duct prototype (inlet).	161
Figure 4.23 New duct prototypes insertion loss.	163
Figure 4.24 Model of the speaker loaded by a duct of length L.....	163
Figure 4.25 Ducts effect on the PC casing transmission loss.....	166
Figure 4.26 Ducts effect on the power supply fan sound power level.	168

Figure 4.27 Ducts effect on the processor fan sound power level.	168
Figure 4.28 Ducts effect on the hard drive sound power level.....	169
Figure 4.29 Overall passive treatment effect on the sound power level of the PC.	171
Figure 5.1 Single channel feedforward controller based on the Filtered-X LMS algorithm	173
Figure 5.2 Secondary source controlling a propagating disturbance in duct.	176
Figure 5.3 Single channel broadband feedforward ANC in duct.....	177
Figure 5.4 Complete block diagram of a SISO feedforward digital controller.....	178
Figure 5.5 Block diagram of ANC system with noise on the input signals	180
Figure 5.6 First duct prototype.	185
Figure 5.7 Single channel ANC in duct. Reference signal type selection.....	187
Figure 5.8 Single channel ANC in duct. Reference signal selection (concluded).	188
Figure 5.9 Reference microphone location effect on the achievable control in duct.	190
Figure 5.10 Microphone windscreen effect on the achievable control in duct.	192
Figure 5.11 Microphone precision effect on the achievable control in duct.	194
Figure 5.12 FRF of a 2 nd order low-pass analog filter.....	197
Figure 5.13 Group delay of a 2 nd order low-pass analog filter.....	197
Figure 5.14 FRF of a 2 nd order high-pass analog filter.	198
Figure 5.15 Group delay of a 2 nd order high-pass analog filter.....	198
Figure 5.16 Filter order effect on the Group Delay. Butterworth, 1000Hz.....	199
Figure 5.17 FRF of various 2 nd order low pass filters, Butterworth type.....	200
Figure 5.18 Cut-on frequency effect on the Group Delay. Butterworth, 2 nd order.	201
Figure 5.19 Effect of microphone-speaker distance on the FRF and Group Delay.	203
Figure 5.20 Effect of the driver size on the FRF and Group Delay (baffled drivers).	204
Figure 5.21 Effect of the acoustical load on the FRF and Group Delay.	205
Figure 5.22 Effect of the acoustical load on the FRF and Group Delay (continued).....	206
Figure 5.23 Noise controller block diagram, causality tests.	209
Figure 5.24 Causality effect on the ANC performance in the first duct prototype.	210
Figure 5.25 Noise controller block diagram, final setup for the first duct prototype.....	212
Figure 5.26 Fans ANC in first duct prototype. Error signal power spectrum.	213
Figure 5.27 Fans ANC in first duct prototype. Error signal attenuation.	213
Figure 5.28 Pictures of the second duct prototype.	215

Figure 5.29 Single channel ANC in inlet duct. Reference signal selection.	218
Figure 5.30 Single channel ANC in outlet duct. Reference signal selection.	219
Figure 5.31 4'' speaker and microphone group for the new ducts prototype.	220
Figure 5.32 Pure tone ANC in new inlet duct prototype. Error microphone level.	222
Figure 5.33 Pure tone ANC in new outlet duct prototype. Error microphone level.	222
Figure 5.34 Broadband ANC in new inlet duct prototype. Error microphone level.	224
Figure 5.35 Broadband ANC in new outlet duct prototype. Error microphone level.	224
Figure 5.36 Block diagram of the 2I2O controller for global duct ANC.	226
Figure 5.37 2I2O ANC in new ducts. All sources on. Inlet duct microphone level.	229
Figure 5.38 2I2O ANC in new ducts. All sources on. Outlet duct microphone level.	230
Figure 5.39 2I2O ANC effect on sound power level. All sources on.	230
Figure 5.40 2I2O ANC effect on sound power level. Power supply fan noise.	231
Figure 5.41 2I2O ANC effect on sound power level. Processor fan noise.	231
Figure 6.1 Block diagram of a local ANC system for controlling noise of PC's.	235
Figure 6.2 Presentation of the PC simulated user environment in anechoic chamber.	239
Figure 6.3 Picture the PC simulated user environment in anechoic chamber.	240
Figure 6.4 SISO local ANC of power supply fan. Reference signal selection.	241
Figure 6.5 SISO local ANC of processor fan. Reference signal selection.	243
Figure 6.6 SISO local ANC of power supply fan. Effect of control filter size.	245
Figure 6.7 SISO local ANC of processor fan. Effect of control filter size.	247
Figure 6.8 Speaker location test setup.	250
Figure 6.9 SISO local ANC of power supply fan. Effect of speaker location.	251
Figure 6.10 Local SISO ANC of power supply fan. Attenuation.	253
Figure 6.11 Local SISO ANC of power supply fan. Error mic power spectrum.	253
Figure 6.12 Local SISO ANC of processor fan. Attenuation.	255
Figure 6.13 Local SISO ANC of processor fan. Error mic power spectrum.	255
Figure 6.14 MISO reference signals selection. Power supply fan noise.	260
Figure 6.15 MISO reference signals selection. Processor fan noise.	263
Figure 6.16 MISO reference signals selection. All sources operating.	264
Figure 6.17 Reference signals effect on MISO ANC. All sources operating.	269
Figure 6.18 MISO Local ANC performance. Effect of primary noise sources complexity.	272

Figure 6.19 Primary and secondary monopole sources, separated by distance d .	274
Figure 6.20 Zone of quiet for an “acoustic doublet” [26].	275
Figure 6.21 Zone of quiet for a “tuned acoustic doublet” [26].	276
Figure 6.22 Picture the MIMO Local ANC setup in the simulated user environment.	280
Figure 6.23 Picture the MIMO Local ANC hardware setup.	281
Figure 6.24 MIMO local ANC setup with the monitoring microphones array.	282
Figure 6.25 Spatial control performance. Power supply fan noise.	284
Figure 6.26 Spatial control performance. Processor fan noise.	285
Figure 6.27 Spatial control performance. Power supply and processor fan noise.	286
Figure 6.28 Spatial control performance. All sources operating.	286
Figure 6.29 Local control performance (left ear error). All sources operating.	288
Figure 6.30 Attenuation of power supply fan BPF (230-235Hz).	289
Figure 6.31 Attenuation of the fourth harmonic of the power supply fan BPF (905-920Hz).	290
Figure 6.32 Attenuation of a high frequency pure tone (1560-1570Hz).	291
Figure 6.33 Spatial control performance with one speaker and one error.	293
Figure 6.34 Spatial control performance with two speakers and one error.	294
Figure 6.35 Spatial control performance with two speakers and two errors.	295

List of Tables

Table 2.1 Exact solution to the MSE minimum in MIMO configuration	73
Table 3.1 Radiated power distribution over the PC casing measured panels.....	114
Table 3.2 Sound power level of the PC sources.....	122
Table 3.3 Total sound pressure level at the head location (user environment).....	124
Table 4.1 Speakers performance comparison table.....	156
Table 5.1 Control setup settings.	210
Table 5.2 SISO controller settings for pure tone control	221
Table 5.3 SISO controller settings for broadband noise control.	223
Table 5.4 2I2O controller settings for broadband noise control.....	227
Table 6.1 SISO local ANC performance. Effect of control filter size.	247
Table 6.2 SISO local ANC performance. Effect of extra parameters.	249
Table 6.3 SISO local ANC performance. Effect of speaker location.....	251
Table 6.4 List of reference signals used in the MIMO local ANC system.	259
Table 6.5 Multiple reference signals selection for the MISO local ANC.	266
Table 6.6 MISO controller settings for the reference signals tests	267
Table 6.7 Reference signals effect on MISO ANC. All sources operating.....	268
Table 6.8 Noise source complexity effect MISO ANC. Configurations tested.	270
Table 6.9 Noise source complexity effect MISO ANC. Results.....	272
Table 6.10 MIMO controller settings for the control path complexity tests.....	292
Table 6.11 Test configurations for the control path tests in MIMO Local ANC.....	292

Chapter 1. Introduction

1.1. Motivation

Heat is becoming one of the most critical issues in computer or semi-conductor design. In early 2001, Intel CTO Pat Gelsinger said it is expected that, in ten years from now, microprocessors will operate at a clock speed of 10 to 20GHz and be capable of processing one trillion operations per second. However, Mr. Gelsinger also suggested that, if no improvements in terms of thermal load were performed, those chips would produce as much heat/mass, for their proportional size, as a nuclear reactor.

Intel co-founder Gordon Moore dictates that the number of transistors on a microprocessor doubles roughly every two years (Moore's law). Over the past 30 years, the increase in transistor density lead to computing advances. However, this miniaturization process was accompanied with an increase in the amount of power required to run a processor, which in turn generates extra heat. Unless insulating techniques are created, the pace of development as predicted by Moore's law will likely slow.

Heat is also the primary source of failure of electronic equipment. Thus, a major effort in the computers design is to ensure proper cooling of the processor units. Small electric fans are typically used for that purpose. Cooling fans become larger in size and/or have faster speed due to the heat increase with modern chips. Often, two to three fans are incorporated in PC units for cooling purposes.

As an indirect consequence of increased heat cooling requirements, computers have thus become noisier and it is expected that the noise annoyance and sound quality of those machines will be unsatisfactory in the near future. Moreover, current data storage devices involve rotating mechanical parts such as hard drives, and also present increased acoustic

noise output due the faster rotating speeds. Lately, hard drive manufacturers have put much effort to reduce the noise of their units. Nevertheless, such components still constitute an extra source of noise in personal computers. Cornell University researchers found that even low-level office noise can increase health risks and lower task motivation for workers [1]. Personal computers are also typically used at home, in rooms where lower background noise levels are present.

Thus, there is a need for finding cost effective solutions to PC noise. An important challenge in personal computers is the size constraint, which makes the implementation of passive noise control techniques very difficult. Consequently, a study of the potential of active noise control (ANC) to develop a cost effective, compact solution to the noise problems discussed above was proposed. This technique seems particularly well suited to computer noise since most of the hardware required for its implementation is present in the computer itself. Moreover, commercial applications of ANC will increase rapidly in the near future as the price of digital signal processors (DSP), analog to digital (A/D), and digital to analog (D/A) converters continues to fall.

1.2. Active Noise Control

Noise control is defined as the technology of obtaining an acceptable noise level at a receiver. Noise control falls into three major categories: (i) noise reduction at the source, (ii) noise control of the transmission path, and (iii) the use of noise protective measures at the receiver.

Until recently, most of the noise control was performed using passive techniques. If we consider silencers, the reactive type aims at reflecting the propagating energy toward the source. On the other hand, in dissipative type silencers, noise reduction can be achieved through dissipation of acoustic energy incurred by viscosity as the sound travels through a porous material. This technique becomes inefficient at low frequency because the thickness of the absorbing material necessary to produce a given attenuation increases with decreasing frequency.

The other noise control technique available, named active noise control, involves production of destructive interference between the unwanted disturbance and an artificially produced sound field. Such technique is referred as “active” because a controllable source of sound is used to perform the noise control.

The French engineer H. Coanda was first to propose the idea of canceling an unwanted sound by destructive interference. His 1930 patent [3] is considered as the first printed document expressing the idea of sound cancellation by destructive interference. Nevertheless, the first observation of acoustic cancellation dates from the end of the 19th century, after J.W. Strutt, Third Baron Rayleigh, published observation concerning the source of difference in tones generated by the superposition of two stopped organ pipes [2] The German physicist P. Lueg patented a scheme to get destructive interference in a duct [4] His approach, illustrated in Figure 1.1, can be referred as feedforward since a microphone, located upstream the control speaker was used to generate a control signal. The electronic controller was adjusted so that loudspeaker produces a canceling wave in the downstream duct. In the mid-fifties, H. F. Olson and E. G. May, of Radio Corporation of America, published the description of an Electronic Sound Absorber, with suggested applications such as sound reduction in vehicles, noise reduction headsets or duct noise reduction [5] Their approach, illustrated in Figure 1.2, is referred as feedback control since the system is based upon detecting the offending sound with a microphone and feeding the signal back through a controller to a control speaker located close to the microphone. W. B. Conover, of General Electric, carried out experiments in the same period concerning the noise reduction of power transformer noise using a loudspeaker located next to the unit as the canceling source [6]

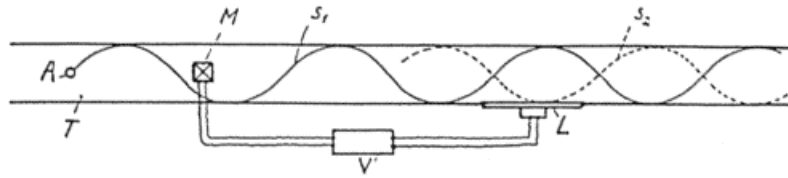


Figure 1.1. Illustration page of Paul Lueg's 1934 patent [4]

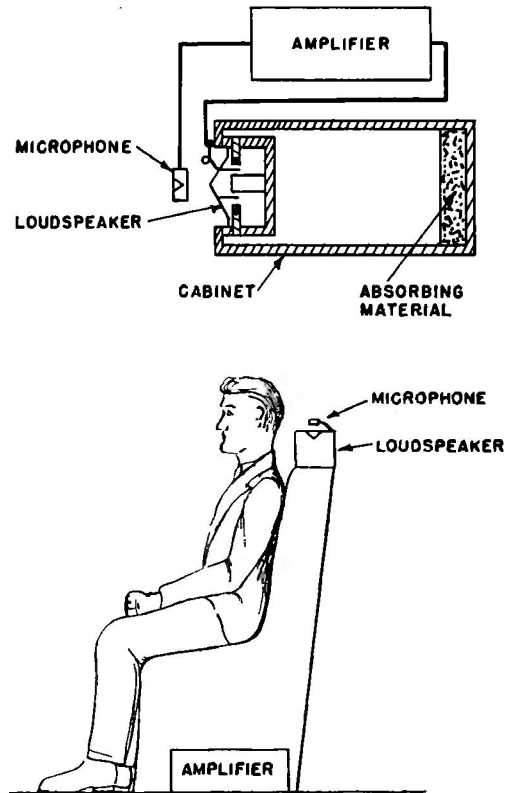


Figure 1.2. Illustration of the work of Olson and May [5].

Although the potential of active noise control had been demonstrated in the 1950s, its practical use was limited for two major reasons. First, for high sound attenuation in the feedforward arrangement, the control signal needs to be adjusted with high accuracy in terms of phase and amplitude, which is difficult to achieve with analog systems. Moreover, for global control of sound, multiple control sources that interact with each other are usually needed. Thus, it was difficult to construct an optimal coupled analog controller that lead to good global performance. In the early 70s, digital signal processing techniques and devices were investigated for active noise control applications by Kido [7] and Chaplin [8]. Their work constitutes the basis of modern feedforward active noise control technique. These control approaches have become easier to implement due to the rapid development of affordable digital devices. Also, some algorithms and techniques were directly developed from the literature of adaptive signal processing.

Researchers, who earlier developed practical active noise control techniques using digital devices, tended to view them as “black boxes” with little knowledge of the system dynamics, particularly when the controller was designed to be adaptive or self-training. Such an approach would lead to good results for simple acoustic systems. Unfortunately, for complex systems, although good attenuation was obtained at the error microphones, poor global control was achieved since the sound field would be often increased at other locations. Thus, the study of physical mechanisms and the associated advantages and limitations of active noise control approach constituted an important area of study in the 90’s. Nelson and Elliott, in particular, have performed an extensive work to study the physics of active noise control [9]. They show that, for good performance, the active sources need to be close to the primary source (with half a wavelength) and the required spacing decreases with frequency. This phenomenon is explained as a combination of the primary and secondary noise sources to achieve a compact higher-order sound source with a lower radiation efficiency.

C.R. Fuller and A.H. Von Flotow summarize the problem and distinguish three main mechanisms of control [10]: (i) local minimization, where an interference field is created with local minima and maxima; (ii) global minimization, in which the power output of the noise source is reduced through a reduction of the radiation resistance; and (iii) power absorption, in which the active sources absorb power from the primary source.

Two main control approaches are used in adaptive active noise control as shown in Figure 1.3. In the case of feedforward control, a reference sensor is used to detect a reference signal (pressure or vibration) that is highly correlated with the disturbance mechanism recorded at the point to be controlled. An error sensor is located in the region where the disturbance needs to be controlled and provides information about the performance of the controller. The reference signal is fed through an adaptive filter (typically a finite impulse response digital filter) whose coefficients are optimized such that the resulting control signal, fed to the secondary source (pressure or vibration actuator), leads to an attenuation of the sound or vibration field at the error sensor. In the case of feedback controller, no information of the disturbance prior to control is used. Feedback controllers, who use the error signal to generate a control signal, are unable to control unpredictable disturbances and tend to be less robust than feedforward controllers. In both cases, the adaptive digital control system is implemented on a digital signal processor (DSP).

The problem of minimization of the error signals using filters is referred as optimal filtering. Although it is possible to fix the design of the control filter, the controller is usually adaptive due to changes in the disturbance characteristics that lead to changes in the plant dynamics. Several algorithms that enable effective implementation of adaptive algorithms in DSP boards have been developed in the past. The most common adaptation law is the Filtered-X Least Mean Square (LMS) algorithm proposed by Burgess [11] and Widrow [12] and will be used for the current project. In this method, the filter weights, which relate the control actuator signal to the recent history of the reference signal, are adapted based on an estimate of the gradient of an error surface. This estimate is based upon the error signal and the filtered reference signal.

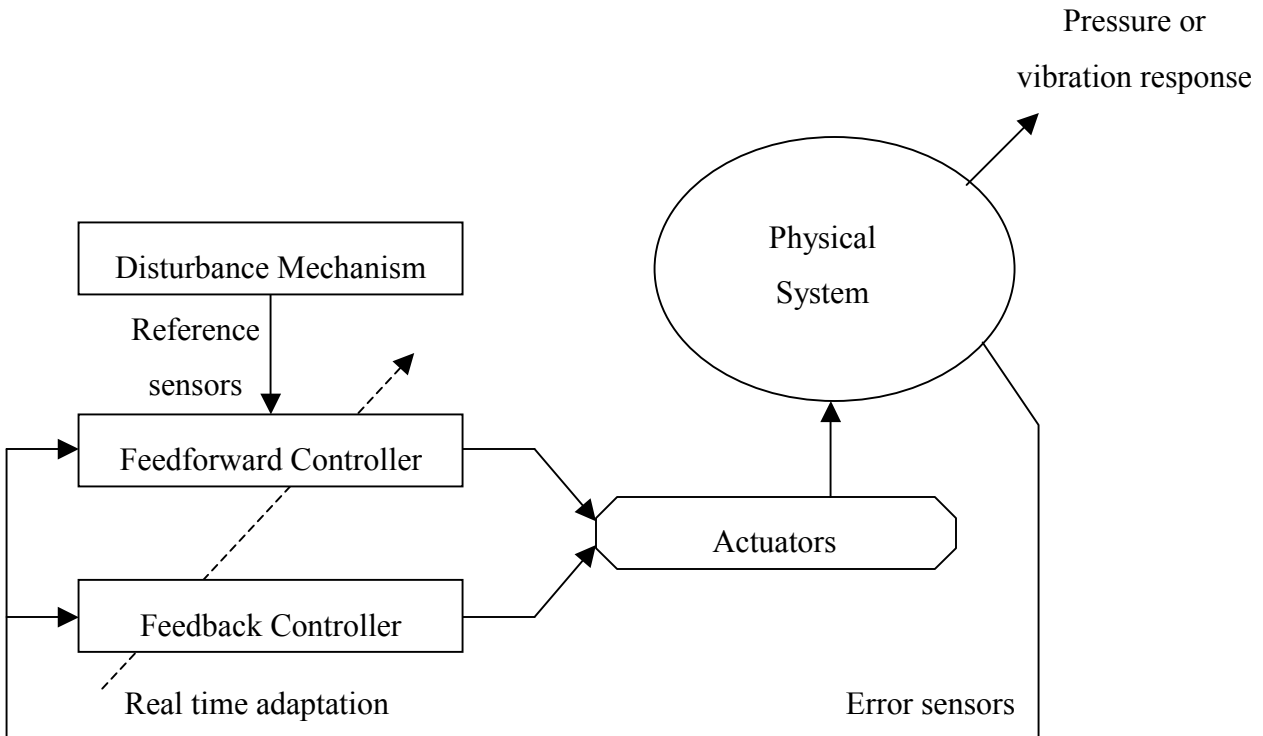


Figure 1.3. Generic control architecture.

Multiple factors can affect ANC performance. Most common issues involve plant modeling errors (due to the finite size of the control filters), causality (when the delay in the control path is larger than in the primary path), coherence (when the disturbance is not perfectly sensed and/or extraneous noise is present), control authority (when the actuators do not couple efficiently to the acoustic or structural medium). Depending on the complexity of the system, each of the issues listed above can be more or less important.

1.3. Objectives and Organization

Very little work has been carried out on the active control of noise from personal computers. On the other hand, most of the techniques developed below such as active noise control in ducts have been extensively covered in the literature. Thus, the objective of this project is to demonstrate the applicability of active noise control techniques to PC noise. The goals of this research can be summarized as follows:

- Identification, characterization and ranking of PC noise sources
- Investigation of active noise control approaches for treating the major PC noise sources
- Establishment of general design rules and trade-off formulae
- Demonstration of the design rules and implementation of the control devices on a suitable PC

Chapter 2 covers fundamentals of digital signal processing, acoustical measurements, and active noise control, and thus constitutes a theoretical basis for all the practical work carried out in this project. The investigation of the Intel PC noise sources is detailed in Chapter 3. Valuable information about the noise sources characteristics and the associated sound radiation paths will help in the design of efficient active and passive control systems. Since active noise control is limited to the low frequency range (typically below 1000Hz), simple passive noise control techniques need to be combined with an active noise control system, which is the subject of Chapter 4. Design rules for proper active noise control system implementation are also detailed in this chapter. The global control of fan noise, based on the design of active / passive sound absorber in ducts, is investigated in Chapter 5. The objective of the global control system is to indeed reduce the PC noise by controlling the sources at the origin. On the other hand, a local active noise control strategy, where the sound field is controlled in a limited space centered around the user head, is investigated in Chapter 6. The advantage of such a strategy is that it does not require a redesign of the computer casing. A cost-effective

implementation can be performed by using the on-board multimedia equipment present in the computer.

Intel Corporation, the sponsor of this project, donated a Dell workstation computer for carrying out experiments and implementing various control strategies.

Chapter 2. Measurements and Control Techniques Review

This chapter constitutes the basis of the work performed during the past two years in the VAL Labs. In any noise problem, the first objective is to accurately characterize the noise sources. This requires knowledge of acoustical measurement standards as will be seen in section 2.2. Also, the acquired data needs to be properly processed in order to provide physically meaningful results. Thus, all the signal processing tools used for various purposes during this project will be introduced and outlined in section 2.1. Then, in order to design an efficient active noise control system, fundamentals of Active Noise Control need to be understood. Section 2.3 constitutes an introduction to adaptive systems and its applications to Active Noise Control.

2.1. Signal Processing Tools

The following algorithms have been used for the processing of all signals included in this thesis. We will refer to [13] and [14] for further explanations of the concepts developed here.

2.1.1. Power Spectrum Estimates

One of the major concerns in the first phase of this project is the determination of the sound power level of various noise sources in the computer. The power spectral density needs to be accurately estimated over a broad frequency range and the overall noise level has to be quantified in a precise manner. For those reasons, specialized signal processing codes were developed, using the Matlab software, in order to process data that was almost exclusively acquired in the time domain. The accuracy of these tools has been

checked by comparing the computed results with those of commercial data-analyzers such as B&K and HP.

2.1.1.1. Discrete Fourier Transform

We start with the expression of the Infinite-Range Fourier Transform of a real-valued continuous signal $x(t)$ defined in [13] by the following:

$$X(f) = \int_{-\infty}^{+\infty} x(t) \cdot e^{-j \cdot 2\pi \cdot f \cdot t} \cdot dt \quad (2.1)$$

For a continuous signal of duration T , the Finite Range Fourier Transform is then expressed as:

$$X(f, T) = \int_0^T x(t) \cdot e^{-j \cdot 2\pi \cdot f \cdot t} \cdot dt \quad (2.2)$$

We assume that the signal $x(t)$ is sampled at N equally spaced points, Δt apart from each other. Due to aliasing problems encountered when manipulating finite duration signals, the acquisition sample rate $F_s = 1/\Delta t = N/T$ is chosen to be at least twice the highest frequency of interest in the studied signal. For more information about the aliasing phenomenon, one should refer to any signal processing textbook such as [13] or [15]. For any frequency f , the discrete version of equation (2.2) is:

$$X(f, T) = \Delta t \cdot \sum_{n=0}^{N-1} x_n \cdot e^{-j \cdot 2\pi \cdot f \cdot n \cdot \Delta t}, \quad (2.3)$$

where: $x_n = x(n \cdot \Delta t) \quad n = 0, 1, 2, \dots, N-1$

Usually, the Discrete Fourier Transform $X(f, T)$ is computed at discrete frequencies f_k , using for example the Fast Fourier Transform algorithm, defined in [13]. We will observe that the computed result of the FFT algorithm is scaled by Δt :

$$X_k = \sum_{n=0}^{N-1} x_n \cdot e^{-j \cdot 2\pi \cdot k \cdot n / N} \quad (2.4)$$

The result of equation (2.4) is computed at the following discrete frequencies:

$$f_k = \frac{k}{T} = \frac{k}{N \cdot \Delta t} \quad k = 0, 1, 2, \dots, N-1$$

The Discrete Inverse Fourier Transform is then defined in [13] as:

$$x_n = \frac{1}{N} \cdot \sum_{k=0}^{N-1} X_k \cdot e^{j2\pi \cdot k \cdot n / N} \quad k = 0, 1, 2, \dots, N-1 \quad (2.5)$$

2.1.1.2. Auto-Spectral Density Function Estimate.

It is assumed that all the signals processed in this project are stationary, which implies that their statistical properties such as the mean square value or standard deviation are not varying in time. Moreover, the random processes studied are considered ergodic so that the time averages, performed over finite time blocks are not time varying either. The discrete time record x_n is divided into n_d segments x_{in} , each of duration T (N elements block). The power spectrums are computed for each sub record, and averaged to obtain an accurate estimate of the power spectral density.

Thus, we define an estimate of the two-sided auto-spectral density function, using the Discrete Fourier Transform:

$$\hat{S}_{xx}(f_k) = \frac{1}{n_d \cdot N \cdot \Delta t} \cdot \sum_{i=1}^{n_d} |X_i(f_k)|^2, \quad (2.6)$$

where:

$$X_i(f_k) = \Delta t \cdot \sum_{n=0}^{N-1} x_{in} \cdot e^{-j2\pi \cdot k \cdot n / N} \quad (2.7)$$

$$f_k = \frac{k}{T} = \frac{k}{N \cdot \Delta t} \quad k = 0, 1, 2, \dots, N-1$$

Since FFT procedures are used to compute the Discrete Fourier Transform of the data-blocks, only the first $N/2$ spectral values are of interest. They correspond to frequencies in the range $0 \leq f_k \leq f_c$ where f_c is defined as the Nyquist frequency. The second half of the spectrum estimate corresponds to negative frequencies. Since the auto-spectra functions are always real valued:

$$\hat{S}_{xx}(f_k) = \hat{S}_{xx}(2f_c - f_k) \quad (2.8)$$

Then, we define the one-sided auto-spectral density function estimate:

$$\hat{G}_{xx}(f_k) = \frac{2}{n_d \cdot N \cdot \Delta t} \cdot \sum_{i=1}^{n_d} |X_i(f_k)|^2 \quad k = 0, 1, 2, \dots, N/2 - 1 \quad (2.9)$$

The result of equation (2.9) has the dimension of power density ($Units^2 / Hz$). Terms can be summed up over the frequency range of interest to estimate the power level of the signal:

$$W = \int_{-\infty}^{+\infty} S_{xx}(f) \cdot df \quad \text{In the continuous domain} \quad (2.10)$$

$$W = \Delta f \cdot \sum_{k=0}^{N/2-1} \hat{G}_{xx}(f_k) \quad \text{In the discrete domain} \quad (2.11)$$

where:

$$\Delta f = \frac{1}{T} = \frac{Fs}{N} = \frac{1}{N \cdot \Delta t} \quad (2.12)$$

The results of equations (2.10) and (2.11) have the dimension of power ($Units^2$). The number of data values N used to compute each Fast Fourier Transform is often called the block size. This parameter determines the resolution of the analysis as can be seen in equation (2.12). Thus, there is a trade-off between resolution of the analysis and number of averages performed, based on a finite length record.

Unfortunately, such procedures induce errors. Indeed, the Finite Fourier Transform of a signal $x(t)$ can be viewed as the Fourier Transform of an unlimited time history record $v(t)$ multiplied by a rectangular time window $u(t)$ of duration T . Thus, the Finite Fourier Transform of $x(t)$ corresponds to the FFT of the original unlimited record $v(t)$ convolved with the FFT of the rectangular window $u(t)$, which is a “damped” sinc function:

$$U(f) = T \cdot \frac{\sin(\pi \cdot f \cdot T)}{\pi \cdot f \cdot T} \cdot e^{-j\pi \cdot f \cdot T} \quad (2.13)$$

Such a spectral window presents a main lobe at $f = 0$ Hz as well as side lobes at other frequencies. The large side lobes of $U(f)$ will allow leakage of power at frequencies well separated from the main lobe of the spectral window and thus will introduce distortions of the estimated spectra. To suppress this leakage problem, we can eventually introduce a time window that tapers the time-history data to eliminate the discontinuities at the beginning and end of the records to be analyzed, thus reducing the amplitude of the side-lobes of the time window. In that case, a compensation factor is included in the Discrete Fourier Transform in order to conserve an accurate estimate of the power level of the original signal. For the Hanning window, it is shown in [13] that the discrete Fourier Transform algorithm becomes:

$$X_i(f_k) = \Delta t \cdot \sqrt{\frac{8}{3}} \cdot \sum_{n=0}^{N-1} x_{in} \cdot \left(1 - \cos^2 \left(\frac{\pi \cdot n}{N} \right) \right) \cdot e^{-j \cdot 2\pi \cdot k \cdot n / N} \quad (2.14)$$

2.1.1.3. Cross-Spectral Density Function Estimate.

The cross-spectral density function estimate can be directly derived from the previous expressions. Thus, we will directly give the definition of interests for our computations. We now consider two discrete time records, $x(t)$ and $y(t)$, divided into n_d segments, each of duration T , corresponding to N samples taken at intervals Δt .

Equation (2.15) represents the single-sided cross-spectral density estimate:

$$\hat{G}_{xy}(f_k) = \frac{2}{n_d \cdot N \cdot \Delta t} \cdot \sum_{i=1}^{n_d} |\bar{X}_i(f_k) \cdot Y_i(f_k)|^2 \quad k = 0, 1, 2, \dots, N/2 - 1 \quad (2.15)$$

where:

$$\bar{X}_i(f_k) = \text{conjugate}(X_i(f_k))$$

$$Y_i(f_k) = \Delta t \cdot \sum_{n=0}^{N-1} y_{in} \cdot e^{-j \cdot 2\pi \cdot k \cdot n / N}$$

2.1.1.4. Accuracy of the Estimate.

It has been proved in [13] that the variance of the power spectral density function estimate is given by the following expressions:

$$\text{Var}[\hat{G}_{xx}(f)] = \frac{1}{n_d} G_{xx}^2(f), \quad (2.16)$$

$$\text{Var}[|\hat{G}_{xy}(f)|] = \frac{1}{\gamma_{xy}^2 \cdot n_d} \cdot |G_{xy}(f)|^2, \quad (2.17)$$

where γ_{xy}^2 is the simple coherence function defined below:

$$\gamma_{xy}^2 = \frac{|G_{xy}|^2}{G_{xx} \cdot G_{yy}} \quad (2.18)$$

Then, we define the Normalized Random Error:

$$\varepsilon[\hat{G}_{xx}(f)] = \frac{1}{\sqrt{n_d}} \quad (2.19)$$

$$\varepsilon[|\hat{G}_{xy}(f)|] = \frac{1}{|\gamma_{xy}| \cdot \sqrt{n_d}} \quad (2.20)$$

From equations (2.19) and (2.20), we deduce that the random error of the previous estimates can be reduced by increasing the number of averages n_d .

2.1.2. Linear SISO Systems Study and Application to ANC

For the following, it is assumed that the studied signals constitute random processes with zero mean value and that we deal with constant-parameter linear systems. A constant-parameter system is time-invariant. A linear system ensures the following properties:

$$f(x_1 + x_2) = f(x_1) + f(x_2) \quad \text{Additive property} \quad (2.21)$$

$$f(c \cdot x) = c \cdot f(x) \quad \text{Homogeneous property} \quad (2.22)$$

2.1.2.1. Dynamic Characteristics

The systems studied here can be characterized by their unit impulse response $h(t)$, which relates any arbitrary input $x(t)$ to the output $y(t)$ by the convolution integral:

$$y(t) = h(t) * x(t) = \int_{-\infty}^{+\infty} h(\tau) \cdot x(t - \tau) \cdot d\tau \quad (2.23)$$

For the system to be physically realizable (causal), we need $h(\tau) = 0$ for $\tau < 0$. Moreover, the constant-parameter linear weighting function $h(\tau)$ has to be absolutely integrable, which insures that the system is stable with a bounded output:

$$\int_{-\infty}^{+\infty} |h(\tau)| \cdot d\tau < \infty \quad \text{for } \tau < 0 \quad (2.24)$$

We can also describe the dynamic of the constant parameters linear system by a frequency response function $H(f)$, which is the Fourier Transform of $h(\tau)$:

$$H(f) = \int_0^{+\infty} h(\tau) \cdot e^{-j \cdot 2\pi \cdot f \cdot \tau} \cdot d\tau \quad (2.25)$$

Letting $X(f)$ and $Y(f)$ be the Fourier Transforms of $x(t)$ and $y(t)$, then a simple expression relates those two quantities to the Frequency Response Function:

$$Y(f) = H(f) \cdot X(f) \quad (2.26)$$

2.1.2.2. Transfer Function Estimate and Simple Coherence Function

In practice, only measured input and output signals that are eventually corrupted by noise are accessible. Thus, the Frequency Response Function needs to be estimated based on these corrupted measurements. Most often, we have to deal with models with uncorrelated output noise.

We consider the following signals:

$$\begin{aligned}x(t) &= u(t), \\y(t) &= v(t) + n(t),\end{aligned}$$

where $u(t)$ and $v(t)$ are considered the true input and output signals, $x(t)$ and $y(t)$ are the measured signals, and $n(t)$ is extraneous noise present on the output signal measurement. For such problems, the frequency response function has to be estimated using the tools previously introduced such as the auto- and cross-spectral density functions estimates:

$$\hat{H}(f_k) = \frac{\hat{G}_{xy}(f_k)}{\hat{G}_{xx}(f_k)} = |\hat{H}(f_k)| \cdot e^{-j\hat{\theta}(f_k)}, \quad (2.27)$$

where:

$$|\hat{H}(f_k)| = \frac{|\hat{G}_{xy}(f_k)|}{\hat{G}_{xx}(f_k)} \quad (2.28)$$

$$\hat{\theta}(f_k) = \tan^{-1} \left[\frac{\text{Im}(\hat{H}(f_k))}{\text{Re}(\hat{H}(f_k))} \right] \quad (2.29)$$

$\hat{G}_{xx}(f_k)$ is an auto-spectral density function estimate of $x(t)$

$\hat{G}_{xy}(f_k)$ is a cross-spectral density function estimate

The spectrum estimates of the input and output measured signals are then related by the following expressions:

$$\hat{G}_{xy}(f_k) = \hat{H}(f_k) \cdot \hat{G}_{xx}(f_k) \quad (2.30)$$

$$\hat{G}_{yy}(f_k) = |\hat{H}(f_k)|^2 \cdot \hat{G}_{xx}(f_k) \quad (2.31)$$

Assuming that $\hat{G}_{xx}(f_k)$ and $\hat{G}_{yy}(f_k)$ are both non-zero, an estimate of the ordinary coherence function between the input $x(t)$ and output $y(t)$ can be deduced:

$$\hat{\gamma}_{xy}^2(f_k) = \frac{|\hat{G}_{xy}(f_k)|^2}{\hat{G}_{xx}(f_k) \cdot \hat{G}_{yy}(f_k)} \quad (2.32)$$

$$0 \leq \hat{\gamma}_{xy}^2(f_k) \leq 1$$

For the ideal case, where there is no uncorrelated output noise, the ordinary coherence function will have a constant value of 1, independent of the frequency. This function will be degraded by the presence of extraneous noise in the measurements, non-linearity, or combination of other inputs effects.

For linear systems, $\hat{\gamma}_{xy}^2(f_k)$ will represent the fractional portion of the mean square value of the output $y(t)$ that is directly correlated with the input $x(t)$ at the frequency f . Indeed, we can then define the coherent output spectrum $\hat{G}_{vv}(f_k)$ and the noise output spectrum $\hat{G}_{nn}(f_k)$:

$$\hat{G}_{vv}(f_k) = \hat{\gamma}_{xy}^2(f_k) \cdot \hat{G}_{yy}(f_k) \quad (2.33)$$

$$\hat{G}_{nn}(f_k) = [1 - \hat{\gamma}_{xy}^2(f_k)] \cdot \hat{G}_{yy}(f_k) \quad (2.34)$$

Here, the ordinary coherence function decomposes the measured output spectrum into uncorrelated components related to either the input signal or the extraneous noise. The simple coherence function will thus be used as a design tool for single input / single output feedforward control strategies where the reference signal needs to be well correlated with the error signal.

The controller is thus considered a linear system. Its input is the reference signal and the output is the error signal. The control strategy is to remove the portion of the error signal that is correlated with the reference signal. Thus, by computing the residual (noise) output spectrum of this particular system, we can predict the most achievable noise attenuation based on the particular reference (input) and error (output) signal combination.

2.1.2.3. Analysis of Discrete Systems

The Fourier Transform defined before is a particular case of the Laplace Transform ($s = 2\pi \cdot j \cdot f$). The Laplace Transform maps the continuous time signals in the s-domain (s is a complex variable) based on the following operation [15]:

$$X(s) = \int_{-\infty}^{+\infty} x(t) \cdot e^{-s \cdot t} \cdot dt \quad (2.35)$$

When introducing the adaptive systems, we will have to deal with discrete data sequences rather than continuous signals. In such a situation, the z-transform, a mathematical tool that has properties similar to the Laplace Transform but applies to discrete sequences only ($z = e^{s \cdot T} = e^{2 \cdot \pi \cdot f \cdot T}$ with T the sampling period), needs to be defined. The infinite range z-transform of a sequence is defined in [15] as:

$$X[z] = \sum_{n=-\infty}^{\infty} x[n] \cdot z^{-n} \quad (2.36)$$

It is also useful to define a FIR filter. Its finite discrete impulse response is constituted by a set of L coefficients w_l that relate the input and output sequences of the filter by a convolution operation:

$$y[n] = \sum_{l=0}^{L-1} w_l \cdot x[n-l] \quad (2.37)$$

If $W[z]$ is defined as the z-transform of the discrete impulse response w , the relationship between the input and output sequences is then more simply expressed in the z-domain by:

$$Y[z] = W[z] \cdot X[z] \quad (2.38)$$

2.1.3. Linear MISO System Study and Application to ANC

2.1.3.1. Traditional Representation, General Relationships.

We consider here q constant-parameter linear systems $H_i(f)$, $i=1,2,\dots,q$, with q measurable inputs $x_i(t)$, $i=1,2,\dots,q$, and one measured output $y(t)$. $X_i(f)$, $i=1,2,\dots,q$, represents the Discrete Fourier Transform of the input record $x_i(t)$. Similarly for $Y(f)$ and $N(f)$ being the transforms of $y(t)$ and $n(t)$:

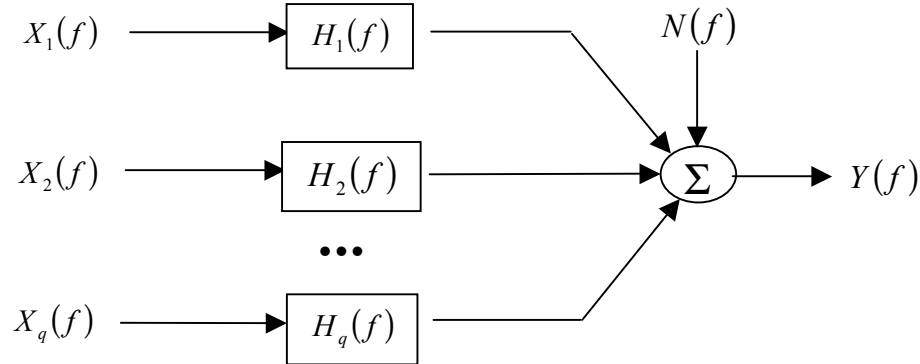


Figure 2.1 Multiple input / single output model for arbitrary inputs

The q inputs can be mutually correlated. Since no real life system is ideal, the output noise term $n(t)$ represents the effects of unmeasured inputs, nonlinear operations, nonstationary effects, and instrument noise. For the model in Figure 2.1 to be well defined, none of the ordinary coherence functions between the inputs should equal unity. If this occurs, one of the inputs does not bring any more information to the model and should be eliminated. Also, none of the ordinary coherence functions between any input

and the output should equal unity. If this happens, the other inputs are not contributing to this output; the model to consider is SISO. Also, the multiple coherence function (defined below) between any input and other inputs should not equal unity. In such a case, the given input is a linear combination of the other inputs and should be removed. Finally, the multiple coherence function between the output and the given inputs should be high enough (above 0.5) for the theoretical assumptions and later conclusions to be reasonable.

We can express the basic relation between $Y(f)$ and $X_i(f)$:

$$Y(f) = \sum_{i=1}^q H_i(f) \cdot X_i(f) + N(f) \quad (2.39)$$

Assuming that the output noise term $n(t)$ is uncorrelated with each $x_i(t)$, we can prove that:

$$G_{iy}(f) = \sum_{j=1}^q H_j(f) \cdot G_{ij}(f) \quad i = 1, 2, \dots, q \quad (2.40)$$

This is a set of q equations and q unknowns, the $H_i(f)$, $i = 1, 2, \dots, q$, where all the spectra are computed from the measured input and output records.

Then, the total output auto-spectral density $G_{yy}(f)$ can be expressed as:

$$G_{yy}(f) = \sum_{i=1}^q \sum_{j=1}^q H_i^*(f) \cdot H_j(f) \cdot G_{ij}(f) + G_{nn}(f) \quad (2.41)$$

If the input records are mutually uncorrelated, much simpler expressions can be deduced:

$$G_{iy}(f) = H_i(f) \cdot G_{ii}(f) \quad i = 1, 2, \dots, q \quad (2.42)$$

$$G_{yy}(f) = \sum_{i=1}^q |H_i(f)|^2 \cdot G_{ii}(f) + G_{nn}(f), \quad (2.43)$$

where:

$$H_i(f) = \frac{G_{iy}(f)}{G_{ii}(f)} \quad i = 1, 2, \dots, q \quad (2.44)$$

Moreover, as a generalization to equations (2.33) and (2.34), the coherent output spectrum $G_{vv}(f)$ and residual output spectrum $G_{nn}(f)$ become in that case:

$$G_{vv}(f_k) = \sum_{i=1}^q \gamma_{iy}^2(f_k) \cdot G_{yy}(f_k) \quad (2.45)$$

$$G_{nn}(f_k) = \left[1 - \sum_{i=1}^q \gamma_{iy}^2(f_k) \right] \cdot G_{yy}(f_k) \quad (2.46)$$

In practice, the measured input records are usually correlated. Then, the previous formulas are not applicable. For example, we consider a double input / single output system with the inputs correlated. The inputs $x_1(t)$ and $x_2(t)$ will lead to the output $y(t)$ through both $H_1(f)$ and $H_2(f)$. Then, the sum of the ordinary coherence functions between the inputs and the output will exceed unity. Thus, the coherent and the residual output power can't be correctly estimated.

In practice, the multiple correlated inputs / single output problem can be solved based on matrix computation or conditioned spectral analysis. During this project, codes were developed based on both models, but only the second technique is detailed below because it is the most intuitive one.

2.1.3.2. Conditioned Input Model.

The arbitrary input model from Figure 2.1 is transformed into a simpler model with an ordered set of conditioned inputs as described in [14]. No change is made to $Y(f)$ and $N(f)$. But, $X_{i-(i-1)}(f)$, $i=1,2,\dots,q$, represents the i th record conditioned on the previous $(i-1)$ records, that is, when the linear effects of $x_1(t)$, $x_2(t)$, up to $x_{i-1}(t)$ have been removed from $x_i(t)$ by optimum linear least squares prediction techniques. These ordered conditioned input records are mutually uncorrelated, contrary to the original arbitrary records. A general rule is to order the input records based on the ordinary coherence function between each input record and the output record.

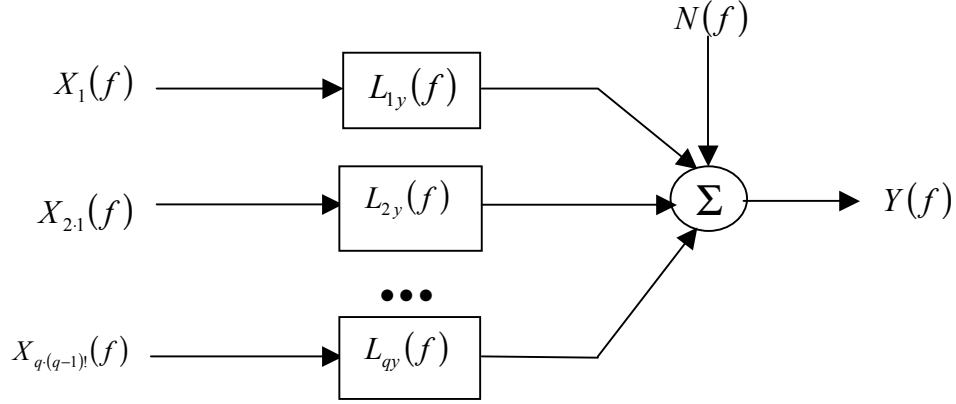


Figure 2.2 Multiple input / single output model for ordered conditioned inputs

a. Spectrum Functions and Notations.

For notation simplifications, we will consider that $y(t) = x_{q+1}(t)$. The auto- and cross-spectral density functions estimates on the input and output records are computed based on the tool developed above. The one-sided spectral density functions will be denoted by:

$$G_{ij}(f_k) = G_{x_i x_j}(f_k) \quad i, j = 1, 2, \dots, q, q+1 \quad (2.47)$$

$$G_{iy}(f_k) = G_{x_i y}(f_k) \quad k = 0, 1, 2, \dots, N/2 - 1 \quad (2.48)$$

The $G_{iy}(f_k)$ terms can be obtained by setting $j = q+1 = y$. An augmented $(q+1) \times (q+1)$ input/output measured spectral density matrix $\{G_{ij}\}$ is obtained. It is explained in reference [14] that this matrix is Hermitian: the terms on the main diagonal are real-valued where as the symmetrical terms across the diagonal are complex conjugates of each other. This property will also be true for the conditioned spectral density matrices defined below.

b. Conditioned Spectral Density and Coherence Functions.

In reference [14], a general algorithm for the computation of conditioned spectral density functions is provided for any $j \geq i$ and $r < j$, where $i = 1, 2, \dots, q, q+1$ and $r = 1, 2, \dots, q$ and at any fixed frequency f_k :

$$G_{ij \cdot r!}(f_k) = G_{ij \cdot (r-1)!}(f_k) - L_{rj}(f_k) \cdot G_{ir \cdot (r-1)!}(f_k), \quad (2.49)$$

where L_{ij} is the optimal frequency response function between records i and j :

$$L_{rj}(f_k) = \frac{G_{rj \cdot (r-1)!}(f_k)}{G_{rr \cdot (r-1)!}(f_k)}. \quad (2.50)$$

The final q^{th} step from the previous equations yields:

$$G_{yy \cdot q!}(f_k) = G_{yy \cdot (q-1)!}(f_k) - L_{qy}(f_k) \cdot G_{yq \cdot (q-1)!}(f_k) \quad (2.51)$$

$$L_{qy}(f_k) = \frac{G_{qy \cdot (q-1)!}(f_k)}{G_{qq \cdot (q-1)!}(f_k)}. \quad (2.52)$$

$G_{yy \cdot q!}$ is the output noise term G_{nn} . It thus can be computed following the precedent iterative process.

Partial coherence functions $\gamma_{ij \cdot (i-1)!}^2$ play the same role as ordinary coherence functions, but are applied to conditioned records instead. They are computed using the following formula, for any $j \geq i$:

$$\gamma_{ij \cdot (i-1)!}^2(f_k) = \frac{|G_{ij \cdot (i-1)!}(f_k)|^2}{G_{ii \cdot (i-1)!}(f_k) \cdot G_{jj \cdot (i-1)!}(f_k)} \quad i = 1, 2, \dots, q, q+1 \quad (2.53)$$

When $j = q+1 = y$:

$$\gamma_{iy \cdot (i-1)!}^2(f_k) = \frac{|G_{iy \cdot (i-1)!}(f_k)|^2}{G_{ii \cdot (i-1)!}(f_k) \cdot G_{yy \cdot (i-1)!}(f_k)} \quad i = 1, 2, \dots, q, q+1 \quad (2.54)$$

By manipulating equations (2.51), (2.52) and (2.53), the noise output spectrum can be expressed in function of output spectrum and the partial coherence function:

$$G_{yy \cdot q!}(f_k) = G_{yy}(f_k) \cdot \prod_{i=1}^q \left(1 - \gamma_{iy \cdot (i-1)!}^2(f_k)\right) \quad (2.55)$$

Multiple coherence functions $\gamma_{j:(j-1)!}^2$ are also an extension of ordinary coherence functions. The multiple coherence function $\gamma_{y:q!}^2$ is the ratio of the ideal output spectrum due to the q measured inputs in the absence of noise to the total output spectrum, which includes noise [14]. The multiple coherence function is related to the partial coherence function:

$$\gamma_{y:q!}^2(f_k) = \frac{G_{y:q!}(f_k)}{G_{yy}(f_k)} = 1 - \left[\frac{G_{yy \cdot q!}(f_k)}{G_{yy}(f_k)} \right] = 1 - \prod_{i=1}^q \left(1 - \gamma_{iy \cdot (i-1)!}^2\right) \quad (2.56)$$

The noise output spectrum is also related to the multiple coherence function:

$$G_{yy \cdot q!}(f_k) = G_{yy}(f_k) \cdot \left(1 - \gamma_{y:q!}^2(f_k)\right) \quad (2.57)$$

A more general expression for the multiple coherence function is given below:

$$\gamma_{j:(j-1)!}^2(f_k) = 1 - \left[\frac{G_{jj \cdot (j-1)!}(f_k)}{G_{jj}(f_k)} \right] \quad j = 1, 2, \dots, q, q+1 \quad (2.58)$$

When $j = q+1 = y$, we obtain the multiple coherence function for the full q -input / single output model as seen earlier.

Conditioned spectral analysis is useful for the prediction of power attenuation in feedforward control strategies involving multiple reference and error sensors. Similarly to what has been developed for the SISO feedforward control strategies with the simple coherence function, the most theoretically achievable noise attenuation can be computed, for each error sensor, using one or several of the reference signals, based on the partial coherence between the studied references and error signal. The noise output spectrum computed in equation (2.55) will represent the measured output spectrum when the MISO controller works perfectly.

2.1.4. Delay Estimation

2.1.4.1. Definitions

a. Phase and Group Velocity.

If we consider a packet of waves of slightly differing frequencies traveling in a medium, each wave will move at a particular speed called the phase velocity. The phase velocity is related to the wavelength λ (and thus to the wavenumber k) and the frequency f :

$$v_\theta = \lambda \cdot f = \frac{2 \cdot \pi \cdot f}{k} \quad (2.59)$$

If this phase velocity is the same for any wavelength, such as for plane acoustic waves in the air, all the waves will propagate in the medium at the same speed. The group velocity can be seen as the overall speed of the packet. In this case, called non-dispersive, the group speed will equal the phase speed and will also be independent of the frequency.

On the other hand, there are cases where the medium is frequency-dispersive. In such case, each wavelength will have an associated phase velocity. Thus, the packet of waves will travel the medium at a group velocity different from the phase velocity. Both phase and group speeds are here frequency dependent. The group velocity also defines the rate at which energy propagates through the medium [29].

b. Group Delay in Linear Systems

Two types of constant parameter linear systems can be distinguished regarding their phase properties [15]. First, systems with linear phase characteristics are considered. In that case, the phase shift at the frequency ω is a linear function of ω :

$$H(j \cdot \omega) = \exp(-j \cdot \omega \cdot t_0) = |H(j \cdot \omega)| \cdot \exp(j \cdot \vartheta(j \cdot \omega)) \quad (2.60)$$

$$|H(j \cdot \omega)| = 1 \quad \vartheta(j \cdot \omega) = -\omega \cdot t_0 \quad (2.61)$$

A system with such frequency response characteristics produces an output that is simply a time shift of the input:

$$y(t) = x(t - t_0) \quad (2.62)$$

In equation (2.62), t_0 represents the Group Delay of the linear system. We verify that the system is non-dispersive since the Group Delay is independent of frequency.

On the other hand, the second type of systems, called frequency dispersive, has non-linear phase characteristics. The phenomenon is illustrated by studying a LTI system with non-linear phase characteristics over a narrow frequency band centered at $\omega = \omega_0$. In this narrow frequency band, the phase response of the LTI system can be approximated by a linear function:

$$H(j \cdot \omega) = \exp(-j \cdot (\theta(\omega_0) + \omega \cdot \alpha(\omega_0))) \quad (2.63)$$

$$|H(j \cdot \omega)| = 1 \quad \vartheta(j \cdot \omega) = -(\theta(\omega_0) + \omega \cdot \alpha(\omega_0)) \quad (2.64)$$

Here the approximate effect of the non-linear phase system on the Fourier Transform of a narrowband input noise (also centered at $\omega = \omega_0$) consists of the multiplication by an overall constant complex factor $\exp(-j \cdot \theta(\omega_0))$ and by a linear phase term $\exp(-j \cdot \omega \cdot \alpha(\omega_0))$ corresponding to a time delay of $\alpha(\omega_0)$ seconds. This time delay represents the Group Delay of the system at the frequency ω_0 .

In one other narrow frequency band centered at the frequency $\omega = \omega_1$, the phase response of the system might be approximated by a different linear function, leading to a different Group Delay $\alpha(\omega_1)$. Thus, the Group Delay for a non-linear phase LTI system is frequency dependent, which implies that it is frequency dispersive.

We will use the previous definitions to estimate the delays in the various paths of our control architectures. Since causality is one of the major limitations in the control of noise in short length ducts, it is necessary to have an idea of the different delays involved in the control system in order to maximize the control efficiency.

2.1.4.2. Group Delay Estimate through Correlation Function Estimate.

The following development applies to auto-correlation functions as well, those being just a particular case where $y = x$. For more details one should refer to reference [15].

a. Cross-Correlation Function Definition.

For stationary random processes $\{x_k(t)\}$ and $\{y_k(t)\}$, the cross-correlation function is defined as:

$$R_{xy}(\tau) = E[x_k(t) \cdot y_k(t + \tau)], \quad (2.65)$$

where E denotes the expected value operator:

$$E[x] = \int_{-\infty}^{+\infty} x \cdot p(x) \cdot dx, \quad (2.66)$$

where $p(x)$ is the probability density function of the variable x .

We now consider two discrete forms of the signals $x(t)$ and $y(t)$, containing N samples at intervals Δt . An unbiased estimate of the sample cross-correlation function can be computed for discrete times lags, r , based on the following definition:

$$\hat{R}_{xy}(r \cdot \Delta t) = \frac{1}{N-r} \cdot \sum_{n=1}^{N-r} x_n \cdot y_{n+r} \quad r = 0, 1, 2, \dots, m \quad m < N \quad (2.67)$$

The computation of the cross-correlation function as defined in equation (2.67) is very intensive. Thus, in practice, the cross-correlation function is usually estimated via Fast Fourier Transform procedures. Originally, the auto- and cross-correlation functions are defined as the inverse Fourier Transform of the auto- and cross-spectral density functions:

$$S_{xy}(f) = \int_{-\infty}^{+\infty} R_{xy}(\tau) \cdot e^{-j2\pi f \cdot \tau} \cdot d\tau \quad (2.68)$$

$$R_{xy}(\tau) = \int_{-\infty}^{+\infty} S_{xy}(f) \cdot e^{j2\pi f \cdot \tau} \cdot df \quad (2.69)$$

Thus the Fast Fourier Transform technique can be used to estimate the auto- and cross-correlation functions in a more efficient way (regarding computational effort) than via direct computation. From the cross-spectral density function estimate applied to the discrete time signals x_n and y_n , we can obtain:

$$\hat{R}_{xy}^s(r \cdot \Delta t) = \sum_{k=0}^{N-1} \hat{S}_{xy}(f_k) \cdot e^{j2\pi k \cdot r / N} \quad r = 0, 1, 2, \dots, N-1 \quad (2.70)$$

The first $N/2$ terms correspond to positive lag values. The second half corresponds to negative lag values. An unbiased estimate of the cross-correlation function becomes:

$$\hat{R}_{xy}(r \cdot \Delta t) = \frac{N/2}{N/2 - r} \hat{R}_{xy}^s(r \cdot \Delta t) \quad r = 0, 1, 2, \dots, N/2 - 1 \quad (2.71)$$

$$\hat{R}_{xy}(-r \cdot \Delta t) = \frac{N/2}{r - N/2} \hat{R}_{xy}^s((N-1) - r) \quad r = 0, 1, 2, \dots, N/2 - 1 \quad (2.72)$$

b. Hilbert Transform and Analytic Signal for Delay Estimation.

The Hilbert transform of a real-valued time domain signal $x(t)$ is another real-valued time domain signal, denoted by $\tilde{x}(t)$, such that $z(t) = x(t) + j \cdot \tilde{x}(t)$ is an analytic signal. From $z(t)$, a magnitude function $A(t)$ and a phase function $\theta(t)$ can be obtained. $A(t)$ represents the envelope of the original function $x(t)$ versus time, where as $\theta(t)$ is the instantaneous phase of $x(t)$ versus time. This tool will thus be used to compute the envelope function of cross-correlation function estimates in order to evaluate the delays in both dispersive and non-dispersive propagation problems.

The Hilbert transform can be defined as a convolution integral:

$$\tilde{x}(t) = \int_{-\infty}^{+\infty} \frac{x(u)}{\pi \cdot (t-u)} \cdot du = x(t) * \left(\frac{1}{\pi} \cdot t \right) \quad (2.73)$$

It can also be defined as a phase shift. Let $\tilde{X}(f)$ be the Fourier Transform of $\tilde{x}(t)$. Then,

$$\tilde{X}(f) = \int_{-\infty}^{+\infty} \tilde{x}(t) \cdot e^{-j \cdot 2\pi \cdot f \cdot t} \cdot dt \quad (2.74)$$

$$\tilde{X}(f) = (-j \cdot \text{sign}(f)) \cdot X(f) \quad (2.75)$$

$$\tilde{X}(f) = |X(f)| \cdot e^{-j[\theta_x(f) + \theta_b(f)]} \quad (2.76)$$

where:

$$\theta_b(f) = +\pi/2 \quad \text{when } f > 0$$

$$\theta_b(f) = -\pi/2 \quad \text{when } f < 0$$

Finally, the Hilbert transform is also defined as the imaginary part of the analytic signal. Let $Z(f)$ be the Fourier Transform of the analytic signal $z(t) = x(t) + j \cdot \tilde{x}(t)$. Then, the inverse Fourier Transform of $Z(f)$ yields:

$$\tilde{x}(t) = \text{Im} \left(\int_{-\infty}^{+\infty} Z(f) \cdot e^{j \cdot 2\pi \cdot f \cdot t} \cdot df \right), \quad (2.77)$$

where:

$$Z(f) = \int_{-\infty}^{+\infty} (x(t) + j \cdot \tilde{x}(t)) \cdot e^{-j \cdot 2\pi \cdot f \cdot t} \cdot dt \quad (2.78)$$

It is shown in reference [15] that:

$$Z(0) = X(0) \quad (2.79)$$

$$Z(f) = 2X(f) \quad \text{when } f > 0 \quad (2.80)$$

$$Z(f) = 0 \quad \text{when } f < 0 \quad (2.81)$$

Thus the analytic function of a signal $x(t)$ can be computed based on its Fourier Transform. This is the most efficient way to compute the Hilbert transform. Similarly, the Hilbert transform of correlation functions can be computed based on power spectral density estimates. Let $R_{z_x z_y}(\tau)$ be the complex-valued cross-correlation function for the following analytic signals:

$$z_x(t) = x(t) + j \cdot \tilde{x}(t) \quad \text{and} \quad z_y(t) = y(t) + j \cdot \tilde{y}(t) \quad (2.82)$$

Assuming that $\tilde{R}_{xy}(\tau)$ is the Hilbert transform of the cross-correlation function $R_{xy}(\tau)$, it can be shown that:

$$R_{z_x z_y}(\tau) = 2 \cdot [R_{xy}(\tau) + j \cdot \tilde{R}_{xy}(\tau)] \quad (2.83)$$

Also, a simple expression relates the Fourier Transform of $R_{z_x z_y}(\tau)$, which is $S_{z_x z_y}(f)$, to the cross-spectral density function of the original signals $x(t)$ and $y(t)$, which is $S_{xy}(f)$:

$$S_{z_x z_y}(f) = 4 \cdot S_{xy}(f) \quad \text{when } f > 0 \quad (2.84)$$

$$S_{z_x z_y}(f) = 0 \quad \text{when } f < 0 \quad (2.85)$$

This yields to the theoretical formulae:

$$R_{xy}(\tau) = \text{Re} \left[2 \cdot \int_0^{+\infty} S_{xy}(f) \cdot e^{j \cdot 2\pi \cdot f \cdot \tau} \cdot df \right] \quad (2.86)$$

$$\tilde{R}_{xy}(\tau) = \text{Im} \left[2 \cdot \int_0^{+\infty} S_{xy}(f) \cdot e^{j \cdot 2\pi \cdot f \cdot \tau} \cdot df \right] \quad (2.87)$$

In the digital domain, the following expressions are obtained for the computation of an unbiased estimate of the complex-valued cross-correlation function via Fourier Transform.

The analysis will be limited to positive lag values:

$$\hat{R}_{xy}(r \cdot \Delta t) = \frac{N/2 \cdot \Delta f}{N/2 - r} \cdot \text{Re} \left[\sum_{k=0}^{N/2-1} \hat{G}_{xy}(k \cdot \Delta f) \cdot e^{j \cdot \pi \cdot k \cdot r / N} \right] \quad (2.88)$$

$$\hat{\tilde{R}}_{xy}(r \cdot \Delta t) = \frac{N/2 \cdot \Delta f}{N/2 - r} \cdot \text{Im} \left[\sum_{k=0}^{N/2-1} \hat{G}_{xy}(k \cdot \Delta f) \cdot e^{j \cdot \pi \cdot k \cdot r / N} \right] \quad (2.89)$$

The squared envelope signal of the cross-correlation function estimate $\hat{R}_{xy}(\tau)$ is given by:

$$\hat{A}_{xy}^2(r \cdot \Delta t) = \hat{R}_{xy}^2(r \cdot \Delta t) + \hat{\tilde{R}}_{xy}^2(r \cdot \Delta t) \quad (2.90)$$

Using equations (2.88), (2.89) and (2.90), the envelope of the cross-correlation function for the input and output signals of a linear SISO system can be estimated. As we said before, the result can be used to estimate the Group Delay induced by the system at any particular frequency. For this purpose, a narrowband white noise whose frequencies is fed to the system, and the cross-correlation function estimate between the narrow-band input and the system response is then computed in order to estimate the group in the narrow frequency band of interest.

For non-dispersive propagation problems, it was observed that the delay in the path is not frequency dependent. Then, any narrowband input signal will provide the necessary information. The Group Delay will here be easily measured as the time at which the cross-correlation function estimate $\hat{R}_{xy}(\tau)$ is maximum.

For dispersive propagation problems, the analytic signal associated with the cross-correlation function estimate is reconstructed in order to accurately evaluate the group and phase delays. The peak of the cross-correlation function estimate will occur at a time delay related to the phase velocity of the propagating waves. On the other hand, the peak of the cross-correlation function envelope will occur at a time delay related to the group velocity of the waves packet.

If such a technique is used, we will have to guess what are the frequencies for which the Group Delay of the system is maximum. This can be difficult in most practical problems. Thus, a more straightforward method, developed in the next section, is useful to determine the Group Delay in any narrow frequency band.

2.1.4.3. Group Delay Estimate through the Frequency Response Function:

Provided that we know the dynamics of the linear system under study, the Group Delay at each frequency can be determined based on the phase of the Frequency Response Function. The Group Delay is related to the slope of the phase response by the following expression [15]:

$$\delta(j \cdot \omega) = -\frac{d(\vartheta(j \cdot \omega))}{d\omega} \quad (2.91)$$

Based on the estimates developed in section 2.1.2.2, the Frequency Response Function of the a given system can be computed by feeding it with a broadband white noise and measuring the response due to that input. Then, the Group Delay can be computed at any frequency by manipulating the phase response. This tool will be used for the local control of fan noise in order to minimize the causality problems due to important lags in the control path compared to the disturbance direct path.

2.2. Acoustical Measurements of Noise Sources

In this project, one of the first goals is to identify the stationary noise sources present in the computer. Certain acoustical measurements that help to characterize the spectral and spatial distribution of acoustic energy radiated by the various sources have to be performed. The measured sources can later be ranked, which helps to determine the major noise radiation paths to cancel.

The tools developed below were used to characterize the various noise sources of the PC. Methods defined by the ISO and/or ANSI standards were applied. For further details about each particular measurement standard, one should refer to references [16] through [22].

2.2.1. Acoustic Power Level

2.2.1.1. Fundamentals

a. Sound Power, Sound Intensity, and Sound Pressure

The sound power level of a source can be calculated from the sound intensity on a closed surface that completely surrounds the source:

$$W = \int_S \vec{I} \cdot \vec{n} \cdot dS, \quad (2.92)$$

where:

W is the sound power in watts (W)

$\vec{I}$ is the sound intensity vector in W/m^2

dS is the element of surface area in square meters (m^2)

$\vec{n}$ is the unity vector normal to the surface area

S is the surface that surrounds the source in m^2 .

In practice, this integral is replaced by a summation; the surface S is divided into N segments, the i -th segment having an area S_i . Equation (2.92) is then approximated by the sum:

$$W = \sum_{i=1}^N I_i \cdot S_i \quad (2.93)$$

Here, I_i is the sound intensity normal to the i -th surface segment and S_i is the area of the i -th surface segment.

Several national and international standards use an approximate relationship between sound pressure and sound intensity to determine the sound power level of the source. In the far field of a source radiating into a free field or free field over a reflecting plane, the sound intensity may be approximated by:

$$I = \frac{p_{rms}^2}{\rho \cdot c}, \quad (2.94)$$

where:

p_{rms}^2 is the mean-square sound pressure at the measuring point in (Pa^2)

ρ is the density of air in kg/m^3

c is the speed of sound in m/s

Then, an expression for the sound power level in decibels relative to 10^{-12} W is deduced:

$$L_w = L_p + 10 \log \frac{S}{S_0} - 10 \log \frac{\rho \cdot c}{400}, \quad (2.95)$$

where L_p is the level of the average mean-square sound pressure on the measurement surface in decibels relative to $2 \cdot 10^{-5} N/m^2$ (Pa). In practice, the last term can usually be neglected for measurements of engineering grade.

b. Reverberant Room Method

The sound power level of a machine can be estimated using several methods. A reverberant environment may be used to determine the sound power level of a noise source from measurements of the space-averaged sound pressure level in a diffuse field. If the source emits broadband noise, ANSI S12.30 [16] or ISO 3741 [17] may be used.

The basic equation used to determine the sound power level using the direct method is:

$$L_w = L_p - 10 \log \frac{T}{T_0} + 10 \log \frac{V}{V_0} + 10 \log \left(1 + \frac{S \cdot \lambda}{8V} \right) - 10 \log \frac{B}{1000} - 14, \quad (2.96)$$

where:

L_p is the space-averaged sound pressure level of the equipment being evaluated in decibels relative to $2\text{E-}5 \text{ Pa}$

T is the reverberation time of the room for the frequency band of interest in seconds ($T_0 = 1 \text{ s}$)

V is the room volume in cubic meter ($V_0 = 1 \text{ m}^3$)

S is the area of all boundary surfaces of the room in m^2

λ is the wavelength corresponding to the center frequency in the octave or one-third-octave band being measured

B is the barometric pressure in millibars (1000 mbars = reference barometric pressure).

c. Precision Methods in Anechoic Chamber

Precision methods for the determination of sound power in anechoic environments are defined by the ANSI S12.35 [18] and ISO 3745 [19] standards. Equation (2.95) is used to determine the sound power level from the sound pressure level. In case of measurements over a reflecting plane, an array of fixed microphone positions is used, the positions being distributed over the surface of a test hemisphere. The microphone positions are usually associated with equal areas of the test hemisphere so that the surface sound pressure level is easily determined:

$$\overline{L_p} = 10 \log \left(\frac{1}{N} \cdot \sum_{i=1}^N 10^{0.1 \cdot L_{pi}} \right), \quad (2.97)$$

where:

$\overline{L_p}$ is the surface sound pressure level, in decibels. Reference: $2\text{E-}5 \text{ Pa}$.

L_{pi} is the band pressure level in the i th area.

N is the number of measurement points (between 10 and 20 usually).

2.2.1.2. Practical Measurements

a. Reverberation Room Method

In a reverberant chamber, the total energy in the room increases until reaching equilibrium. At the equilibrium, the amount of energy absorbed by the room equals the amount of energy generated by the source. By measuring the steady state sound pressure level in these conditions, the sound power level of the source can be determined. In theory, the sound pressure is constant in the room, since the sound is completely diffuse. In practice, the sound pressure level is measured at different locations (at least 8) in order to perform a spatial average of the results obtained. The Dell computer in a VAL reverberation chamber with two B&K microphones is shown in Figure 2.3. The major advantages of this measurement method are a simple test setup (as compared to free field method), a precise measurement (as opposed to the free field where the sound pressure is measured at few discrete radiation angles), and the lack of need for sensitive sensors (since the sound pressure levels are much higher than in free field). The largest disadvantage is that no information on source directionality can be inferred from such measurement method (as opposed to the free field method).



Figure 2.3 Sound power level measurement in one of the VAL reverberation chambers

b. Anechoic Chamber Method

The second method for measuring the sound power level requires the use of an anechoic chamber. In this case, the computer is placed over a reflecting surface. A hemisphere whose radius is at least 1 meter is used to dispose microphones at strategic locations. These microphones (10 is a minimum) are located such that they represent an equal area of the surface of hemisphere surrounding the machine. It is assumed that these microphones are in the far field of the radiation: $kd \gg 1$ with k the wavenumber and d the distance to the source. The pressure level squared, which is then proportional to the intensity level, is multiplied by the associated area segment and summed over the hemisphere surface in order to deduce the source power level. The major advantage of such method is that directionality information can be inferred from these measurements. On the other hand, due to the discretization of the sensing field, this method is approximate and can lead to biased results for sources with strong directionality.

The Dell computer in a VAL anechoic chamber is shown in Figure 2.4. We can observe a set of B&K microphones placed at several locations on a hemisphere whose radius is 1 meter.

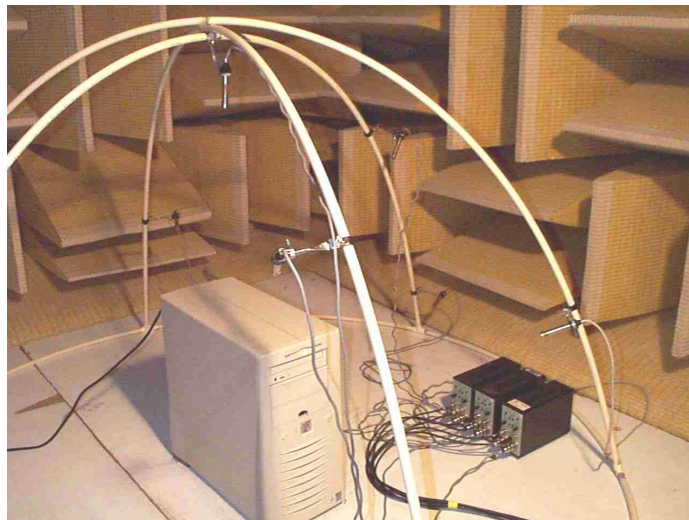


Figure 2.4 Sound power level measurement in one of the VAL anechoic chambers

2.2.2. Intensity Survey

Sound intensity is a measure of the magnitude and direction of the flow of sound energy. The intensity of a sound field is defined as the averaged sound energy transmitted per unit of time, through a unit area, in a specified direction normal to this surface. The sound power generated by a source must be equal to the normal component of the sound intensity integrated over any surface that completely encloses the source.

Thus, sound intensity measurements are very useful for determining the sound power of large machinery in-situ. Nevertheless, the main objective of this series of experiments is to study the radiation of the noise sources through the PC casing. We will focus on the energy flow distribution over the casing, looking for the radiation “hot spots” to indicate the eventual work to be done on treating the casing in order to reduce the overall noise level.

For a detailed study of sound intensity theory, applications, and limitations, one should refer to the book of F.J. Fahy, *Sound Intensity* [20]. References [21] and [22] provide information about standard procedures to measure intensity and their use for sound power level determination.

2.2.2.1. Theoretical Background.

a. Introduction to the Two-microphone Measurement Technique

We start with a basic expression for the instantaneous sound intensity in a particular direction of the particle velocity $u(t)$ considering it is in phase with the pressure $p(t)$:

$$I_{inst}(t) = p(t) \cdot u_r(t), \quad (2.98)$$

where:

$p(t)$ is the sound pressure in Pa

$u_r(t)$ is the particle velocity in the direction of interest in $\frac{m}{s}$.

Choosing the time average T sufficiently large, the net rate of flow of acoustic energy per unit area corresponding to the time-averaged intensity I_r can be deduced:

$$I_r = \frac{1}{T} \cdot \int_0^T p(t) \cdot u_r(t) \cdot dt \quad (2.99)$$

To calculate the intensity, $p(t)$ and $u_r(t)$ are to be determined. The sound pressure $p(t)$ can be easily measured using a microphone. It is more difficult to measure the particle velocity $u_r(t)$. In practice, $u_r(t)$ is estimated by two closely spaced microphones. The idea is to determine the pressure gradient since it is related to the particle acceleration by Euler's equation [26].

Equation (2.100) is a simplified expression of Euler's equation, only one spatial direction r is considered:

$$\frac{\partial p(t)}{\partial r} + \rho \frac{\partial u_r(t)}{\partial t} = 0 \quad (2.100)$$

From equation (2.100), the particle velocity can be deduced in the r direction by approximating the pressure gradient and integrating the differential equation in the time domain:

$$\hat{u}_r(t) = -\frac{1}{\rho} \int_{-\infty}^t \frac{p_2(\tau) - p_1(\tau)}{\Delta r} d\tau, \quad (2.101)$$

where:

Δr is the microphones separation distance in m

$p_i(t)$ is the sound pressure at microphone location i

ρ is the density of air in kg/m^3

The sound pressure at the center of the probe is estimated by:

$$\hat{p}(t) = \frac{p_1(t) + p_2(t)}{2} \quad (2.102)$$

Finally, the results from equations (2.101) and (2.102) have to be integrated in equation (2.99) in order to obtain the time-averaged intensity in the r direction at the 2-microphones probe location. Nevertheless, we have to observe that the relative phase of particle velocity and pressure level was not taken into account to simplify the analysis.

b. Intensity Calculation through the Cross-Spectral Density Estimate

In a general sound field, the particle velocity and pressure are not in phase. Considering a simple harmonic sound field, equation (2.99) can be expressed in a form taking into account the phase information:

$$I_r(t) = \frac{1}{T} \cdot \int_0^T p \cdot \cos(\omega \cdot t + \vartheta) \cdot u_r \cdot \cos(\omega \cdot t + \gamma) \cdot dt, \quad (2.103)$$

where ϑ and γ are arbitrary phase angles. The previous integral can be expressed in terms of active and reactive parts by writing the pressure and particle velocity in complex form. The sound intensity can then be simply express, for a single frequency, in the r direction:

$$I_r(t) = \frac{1}{2} \cdot \text{Re} \left\{ p(t) \cdot \overline{u_r(t)} \right\}, \quad (2.104)$$

where both the sound pressure $p(t)$ and the particle velocity $\overline{u_r(t)}$ are complex exponential quantities, and $\overline{u_r(t)}$ denotes the complex conjugate of $u_r(t)$.

Equation (2.104) can be expressed in the frequency domain:

$$I_r(\omega) = \frac{1}{2} \cdot \text{Re} \left\{ P(\omega) \cdot \overline{U_r(\omega)} \right\}, \quad (2.105)$$

where $P(\omega)$ and $\overline{U_r}(\omega)$ are the Fourier Transforms of $p(t)$ and $\overline{u_r}(t)$. As we have seen in the previous section, the pressure as well as the particle velocity in the direction of interest need to be estimated.

Equation (2.100) can be expressed in the frequency domain by applying the infinite-range Fourier Transform to each member and also approximating the gradient by a finite difference:

$$\int_{-\infty}^{+\infty} \left(\frac{\partial u_r(t)}{\partial t} \cdot \exp(-j \cdot \omega \cdot t) \cdot dt \right) = -\frac{1}{\rho \cdot \Delta r} \cdot [P_2(\omega) - P_1(\omega)] \quad (2.106)$$

After integration, equation (2.106) becomes:

$$U_r(\omega) = \frac{j}{\omega \cdot \rho \cdot \Delta r} \cdot [P_2(\omega) - P_1(\omega)] \quad (2.107)$$

The frequency domain version of equation (2.102) is given by:

$$\hat{P}(\omega) = \frac{P_1(\omega) + P_2(\omega)}{2} \quad (2.108)$$

By inserting the results of equations (2.107) and (2.108) in equation (2.105), we obtain the frequency domain formulation of the sound intensity in the r direction:

$$\hat{I}_r(\omega) = -\frac{1}{2 \cdot \omega \cdot \rho \cdot \Delta r} \cdot \text{Im} \left\{ \overline{P}_1(\omega) \cdot P_2(\omega) \right\} \quad (2.109)$$

Using the cross-spectral density function estimate between the two microphone signals, equation (2.109) can be written in a different form:

$$\hat{I}_r(\omega) = -\frac{1}{\omega \cdot \rho \cdot \Delta r} \cdot \text{Im} \left\{ \hat{G}_{12}(\omega) \right\}, \quad (2.110)$$

where $\hat{G}_{12}(\omega)$ is the single-sided cross-spectral density function estimate between the 2 microphone signals.

2.2.2.2. Accuracy of the Estimate

The frequency domain formulation of the sound intensity developed in section 2.2.2.1 is the method the most widely used in practice. One limitation is the frequency range since the finite difference estimate of the sound intensity is accurate only if the distance between the microphones is much less than the wavelength. However, if too close, phase differences between the microphones will be very small at low frequencies (where the wavelength is very long), thus leading to biased results if the sensors are not perfectly phase matched. Thus, there is a frequency limit depending upon the separation distance of the two microphones.

The finite difference error, that is, the ratio of the measured sound intensity $\hat{I}_r$ to the true intensity I_r , is shown in [20] to be:

$$\frac{\hat{I}_r(\omega)}{I_r(\omega)} = \frac{\sin(k \cdot \Delta r)}{k \cdot \Delta r}, \quad (2.111)$$

where:

$\hat{I}_r(\omega)$ is the biased intensity level

$I_r(\omega)$ is the true intensity level

$k = \omega/c$ is the wavenumber at the frequency of interest ω

In practice, a spacer of 12mm was used that ensures accurate intensity measurement up to 5kHz. One other important aspect is the phase mismatch between the two measurement channels. If a phase mismatch of α_e degrees between the microphones is assumed, a good approximation of the error on the intensity measurement is given by:

$$\hat{I}_r(\omega) = I_r(\omega) - \frac{\alpha_e}{k \cdot \Delta r} \cdot \frac{P_{rms}^2}{\rho \cdot c}, \quad (2.112)$$

where:

α_e is the phase mismatch between the microphones at the frequency of interest (degrees)

P_{rms}^2 is the mean squared averaged pressure at the microphones

As can be seen in equation (2.112), the bias error in the measured intensity is proportional to the phase error and to the mean square pressure. Standard measurement procedures include performance evaluation tests that ensure that the phase error is reasonable. In our experiments, B&K intensity probe with a phase-matched pair of microphones was used.

2.2.2.3. Practical Applications

Intensity measurements can be used for several purposes. One of them is the estimate of the Sound Power Level of a noise source, on site, when regular power level determination techniques are inapplicable. In that case, one has to perform a scan over a surface surrounding the source. This intensity component that is normal to the scanned surface is measured at discrete locations, each associated with a surface similar to the sound power level measurement technique in anechoic environment (i.e. with microphones on an hemisphere). The intensity level is then integrated over the surface in order to deduce the sound power level of the radiating source (refer to section 2.2.1 for more details). Nevertheless, it has been observed that better results can be achieved by performing a single data acquisition, moving the probe along the surface [20].

The second application of sound intensity is the survey of complex noise sources to identify and rank the noise sources and transmission paths. Sound intensity scanning of a system helps to determine the acoustic radiation “hot spots” of the structure. In that case, the intensity level needs to be measured at discrete locations. The results are then used to generate a 3-dimensional plot to get an idea of the major sources of noise. For this project, an automated traverse was used in order to scan the PC surface as can be seen in Figure 2.5.

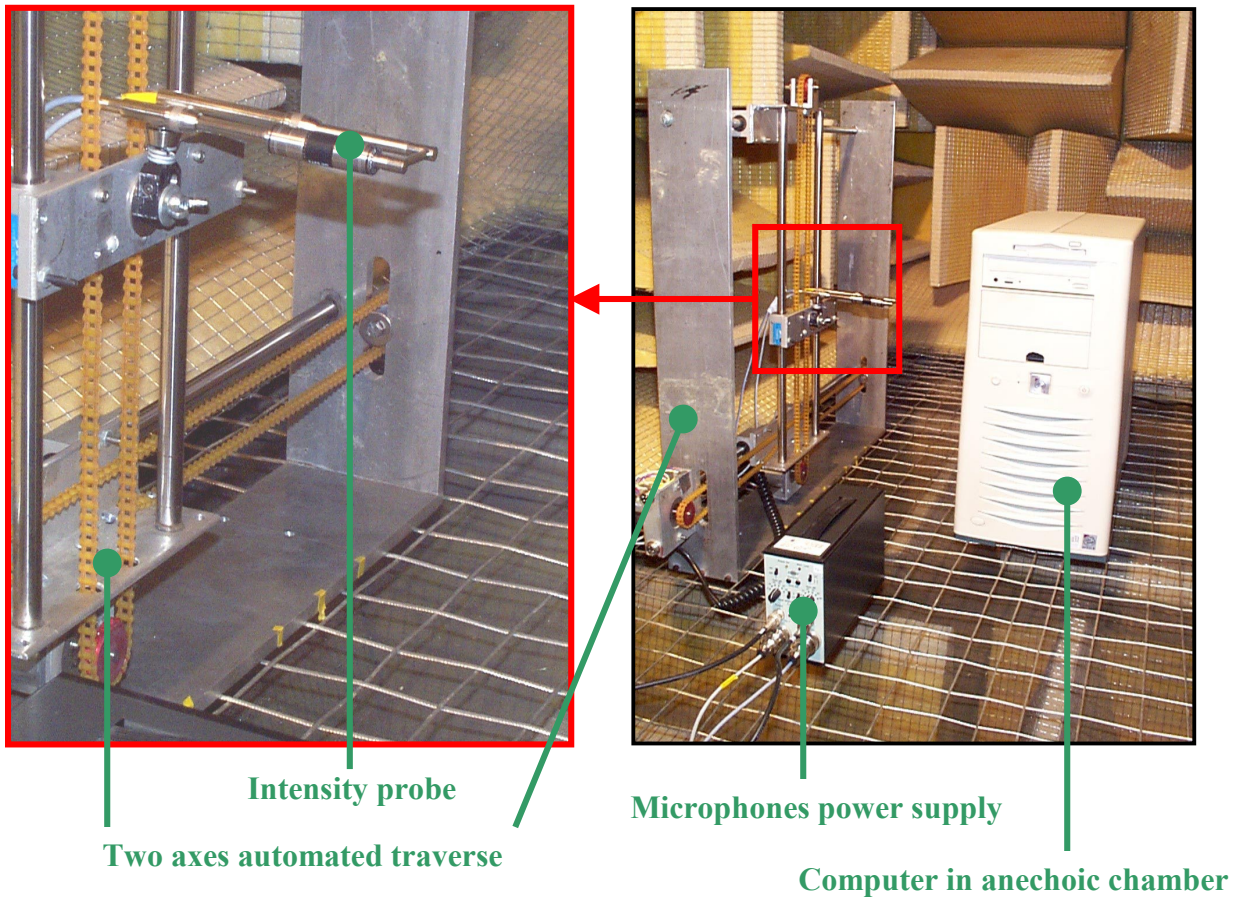


Figure 2.5 Sound power level measurement in one of the VAL anechoic chambers

2.3. Active Noise Control

Acoustic noise and vibration problems are common in the industry, the defense, or even in public spaces as well as homes. Sources of noise in the industry can typically be engines, fans, blowers, transformers, or compressors. To a lesser extent, many household appliances such as dishwashers, dryers, washing machines or vacuum cleaners, can be noisy.

Traditionally, acoustic noise control uses passives techniques such as enclosures, sound barriers and silencers [25]. Such passive techniques are typically used for mid to high-frequency problems (above 1000Hz). Indeed, we can take the example of the sound barrier. Below the critical frequency (the frequency at which the acoustic and structural bending wavelengths match, thus leading to efficient sound radiation by the panel), the sound transmission through panels is dominated by what is referred as the mass law [27]. In practice, the lower the frequency, the heavier and larger will be the barrier in order to be efficient. Another example concerns dissipative silencers, based on energy losses when the sound field is traveling through mediums such as foam and fibers. It will be shown in this thesis that the silencer dimensions are closely related to the wavelength of the sound field to control, making its use impractical at lower frequencies, where the acoustic wavelength is large.

Thus, active sound and vibration control has found its use in the 50-1000Hz frequency range where passive methods are prohibitive or impractical. In Active Noise Control, rapidly changing disturbances are of less importance than “relatively” steady state disturbances. Moreover, many of the common sound and vibration problems are produced by rotating machinery. In such case, the resulting noise is usually characterized by multiple steady state pure tones as well as stationary random noise. This observation will be later verified in the case of PC noise.

Nelson and Elliot [9] have explained in detail the physics of Active Noise Control. It is based upon destructive interference of sound and vibration fields. This principle assumes that both the noise and control systems are linear since the principle of superposition is applied. The main idea is to generate a sound or vibration field that will destructively interfere with the primary disturbance in order to create an overall attenuation of the sound pressure or vibration level.

A generic control architecture is shown in Figure 2.6. Two control approaches can be distinguished. One is the *feedforward* control system where a knowledge of the disturbance, prior to control, is used to generate a control signal. The other approach, a *feedback* controller arrangement, bases exclusively on the sensing of the sound pressure or vibration field to control. Regardless of the controller type, actuators such as speakers (pressure actuator) or piezo-transducers and shakers (vibration actuators) are used to generate a control signal in an attempt to reduce the pressure or vibration level at the error sensor location.

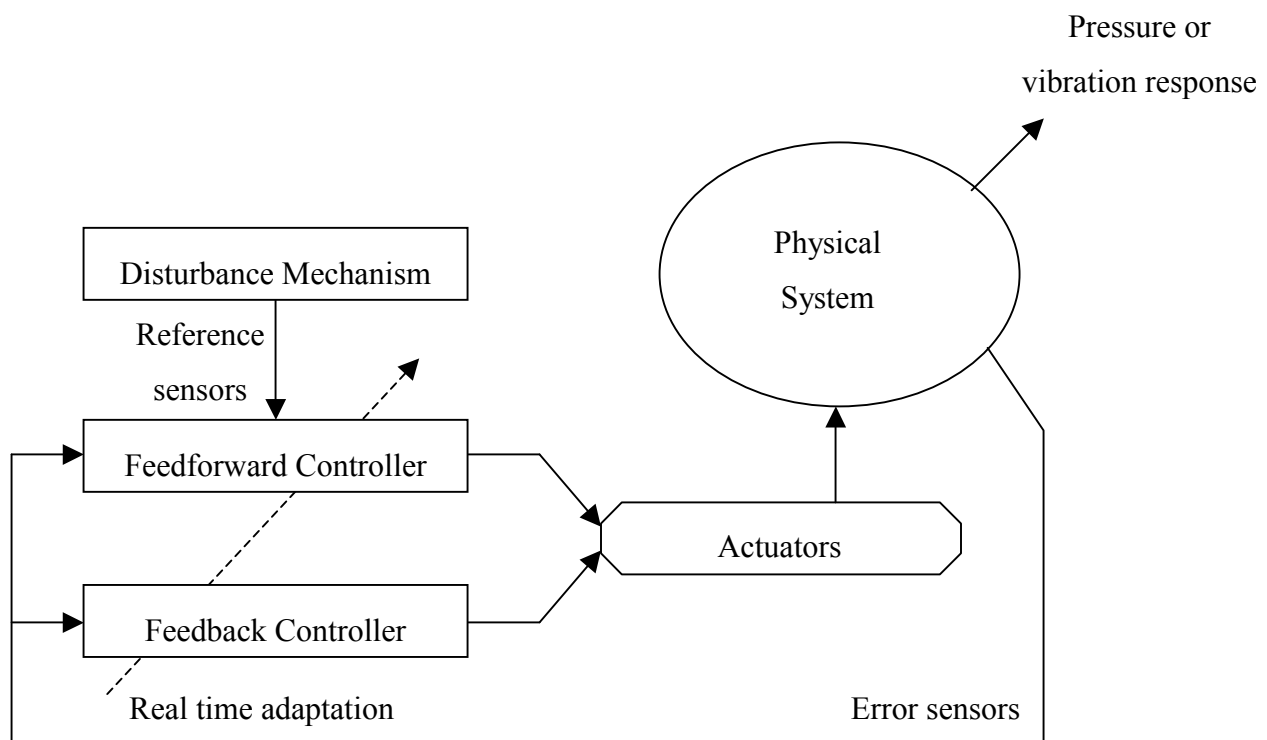


Figure 2.6 Generic control architecture

The error sensor is typically a microphone (pressure sensor) or an accelerometer (vibration sensor). In the case of a feedforward controller, sensors are also used to obtain an early knowledge of the disturbance. Such sensors are usually of the same type as the error sensors.

In this project, only feedforward type controllers will be investigated since they address well the problems we try to solve. The problem of feedback controllers is that they cannot perform attenuation of unpredictable disturbance due to their acausal operation mode (the disturbance needs to be sensed prior to control in order to compensate for the inherent delay in the control path). Moreover, feedback controllers tend to be more sensitive to instability than feedforward controllers. Stability can ultimately be insured by limiting the feedback controller performance (by increasing the phase and gain margin) [9].

In the case of a feedforward controller, the reference signal is appropriately filtered such that the resulting control signal will lead to a minimization of the error signal, as shown in Figure 2.7.

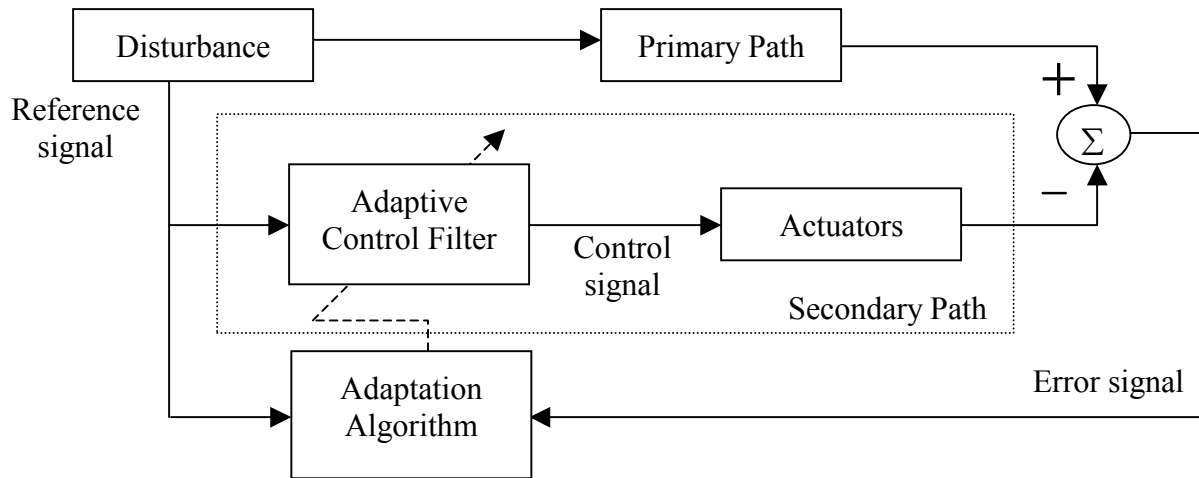


Figure 2.7 Adaptive feedforward controller architecture

In this chapter, adaptive systems, which are the base of Active Noise Control, will be introduced. Also, the noise control strategy used for our noise problems will be developed. For further details about Active Noise Control, one should refer to the many textbooks available on the subject, such as [9], [23], and [24].

2.3.1. Adaptive Systems

Many real life systems are not time invariant. Then, even if some controlling filter is designed with optimal coefficients, the results will not necessarily be optimal since the system characteristics may have changed after the optimization process. Thus, the control filters sometimes need to be adjusted due to plant and / or disturbance variations. The coefficients of the adaptive filter are thus continuously updated, using an adaptive algorithm, in order to track such changes.

2.3.1.1. Digital Filters

Digital filters represent practical tools to design an adaptive controller for many reasons such as flexibility, adaptability, accuracy, stability, and cost. Those filters treat discrete signals by performing a constant set of linear operations on the input data sequence, $x[n]$, in order to create a corresponding output sequence, $y[n]$, based on the value of their coefficients.

Two fundamental types of digital filters can be identified as clearly explained in [15]. A generic block diagram of a digital filter is shown in Figure 2.8. The first type of filter is called Finite Impulse Response (FIR). Such a filter has no feedback loop, the output sequence being exclusively composed of delayed input values ($a_i = 0$ for $i = 1:na$ and $a_0 = 1$). FIR filters are always stable [15] and provide a linear phase response since they contain no poles. The second type of digital filter is called Infinite Impulse Response (IIR). It incorporates both zeros and poles since the output sequence is also composed of delayed output values. In such case, stability is not guaranteed because of the presence of poles. In this chapter, we will look exclusively at FIR filters since they are the only ones used in this research.

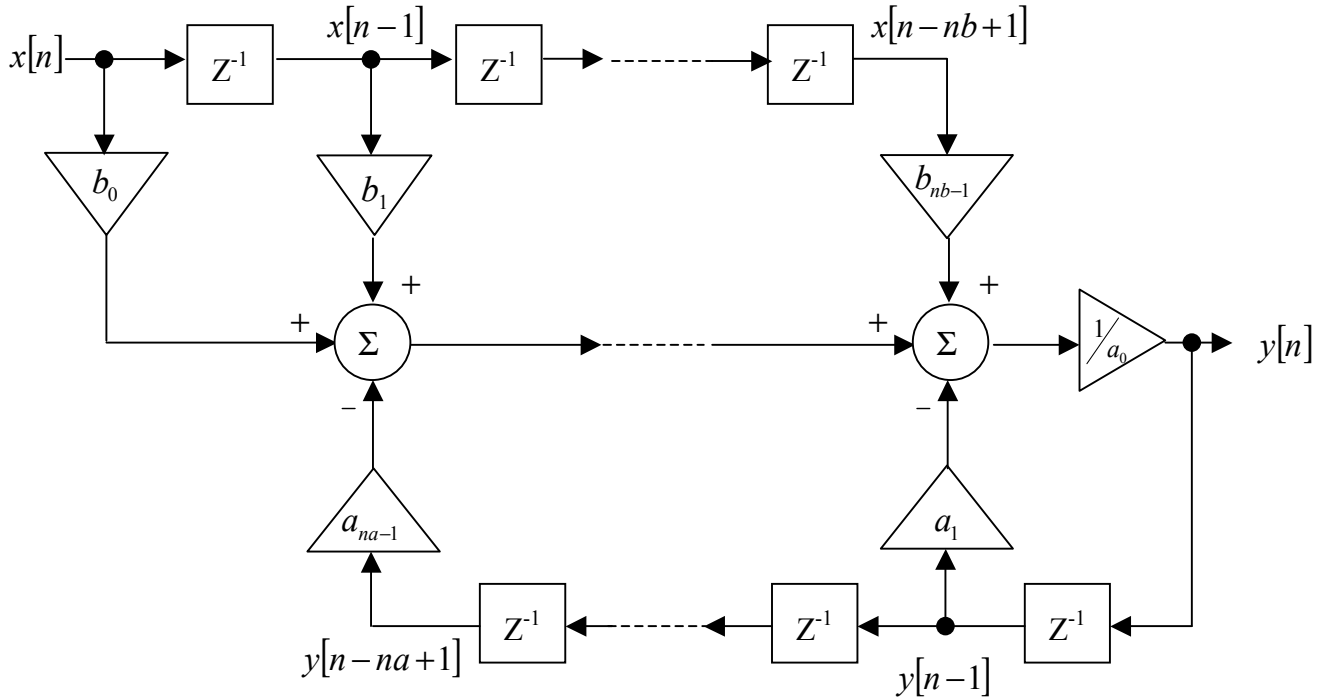


Figure 2.8 Generic block diagram of a digital filter

The impulse response of a FIR filter goes to zero after a finite number of samples. The output sequence $y[n]$ of such a filter can be calculated as follows:

$$y[n] = \sum_{l=0}^{L-1} w_l \cdot x[n-l], \quad (2.113)$$

where w_l is the l^{th} coefficient of the L coefficients filter, and $x[n-l]$ is an element of the input data sequence delayed of l samples. A block diagram representing the implementation of such a filter is shown in Figure 2.8.

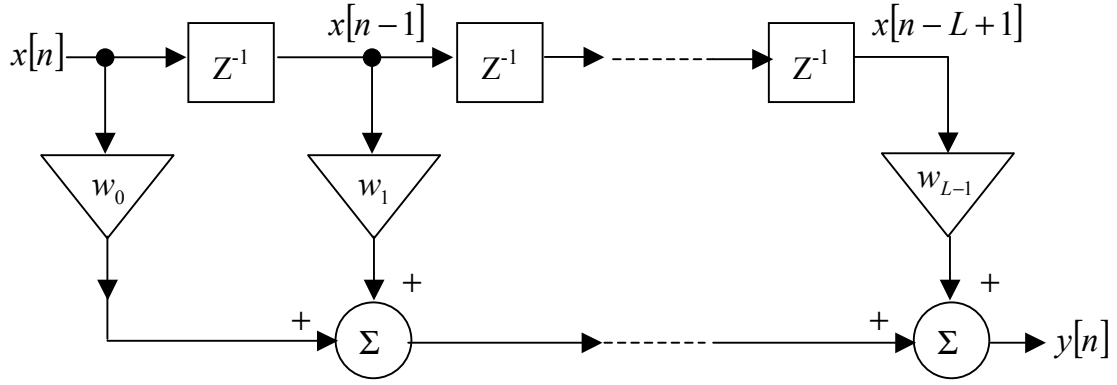


Figure 2.9 Block diagram of an L coefficients FIR digital filter

Let define the z-transforms $X[z]$, $W[z]$ and $Y[z]$ of the sequences $x[n]$, $\{w_0 \ w_1 \ \dots \ w_{L-1}\}$ and $y[n]$:

$$X[z] = \sum_{n=-\infty}^{\infty} x[n] \cdot z^{-n} \quad (2.114)$$

$$W[z] = \sum_{l=0}^{L-1} w_l \cdot z^{-l} \quad (2.115)$$

$$Y[z] = \sum_{n=-\infty}^{\infty} y[n] \cdot z^{-n} \quad (2.116)$$

Then, as we have seen in the previous section, the relationship between input and output sequence of the filter is more simply expressed in the z-domain:

$$Y[z] = W[z] \cdot X[z] \quad (2.117)$$

2.3.1.2. Adaptive Filter

An adaptive filter is composed of a digital filter and an adaptive algorithm adjusting the filter weights so that a desired response is achieved. A general form of adaptive filter is shown in Figure 2.10.

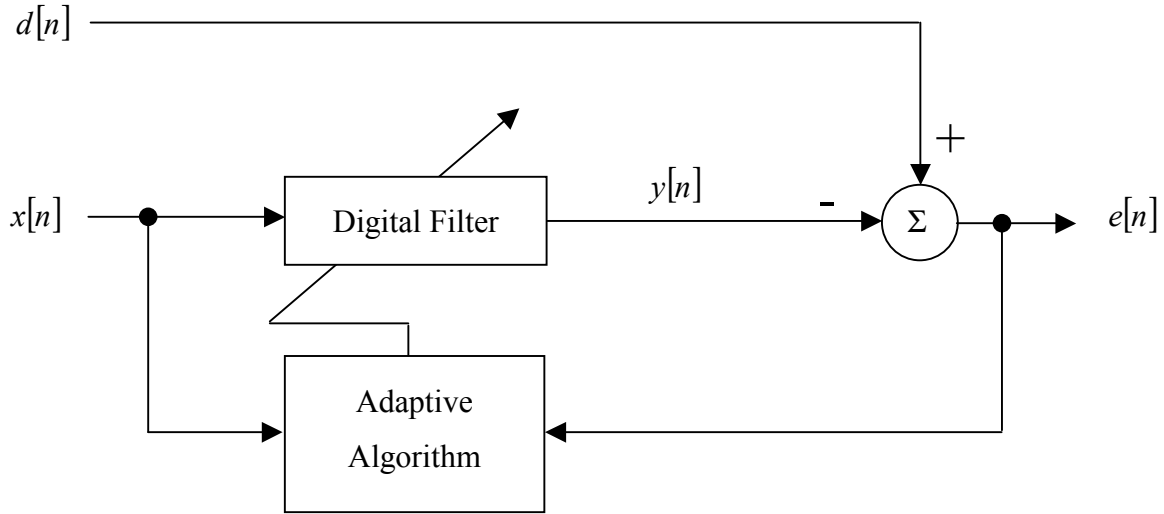


Figure 2.10 Block Diagram of an adaptive filter

In Figure 2.10, $d[n]$ is a desired response or a disturbance signal coming from the primary source, and $y[n]$ is the output of the programmable digital filter. The input sequence to that filter is the reference signal $x[n]$. The error signal $e[n]$ is the difference between the controller output and the disturbance. The adaptive algorithm is used to adjust the coefficients of the programmable filter in order to minimize the mean-square value of the error signal $e[n]$.

The block diagram of the adaptive FIR filter is similar to Figure 2.9 except that the filter weights w_l are replaced by time dependent weights $w_l[n]$ since the filter is continuously updated, sample after sample. The output sequence of the filter is now expressed:

$$y[n] = \sum_{l=0}^{L-1} w_l[n] \cdot x[n-l] \quad (2.118)$$

To simplify the notation, a vectorial notation for the input signal and filter weights is defined [24]:

$$\mathbf{x}[n] \equiv \{x[n] \quad x[n-1] \quad \dots \quad x[n-L+1]\}^T \quad (2.119)$$

$$\mathbf{w}[n] \equiv \{w_0[n] \quad w_1[n] \quad \dots \quad w_n[n]\}^T \quad (2.120)$$

Then, the output sequence can be expressed as the simple product of those last 2 vectors:

$$y[n] = \mathbf{w}[n]^T \cdot \mathbf{x}[n] = \mathbf{x}[n]^T \cdot \mathbf{w}[n] \quad (2.121)$$

Also the error signal becomes:

$$e[n] = d[n] - \mathbf{w}[n]^T \cdot \mathbf{x}[n] \quad (2.122)$$

2.3.1.3. Least Mean Square Adaptive Algorithm

We have seen in the previous section that an adaptive algorithm is used to update the weight vector defining the programmable filter so that the error sequence is minimized. In general, we try to minimize the mean square value of the error signal. Different algorithms have been developed but we will limit our analysis to the Least Mean Square algorithm since it is the one used in the research discussed in this thesis.

a. Mean Square Error

The Mean Square Error (MSE) is defined in [9] as the expectation value of the squared error signal:

$$\xi[n] = E[e^2[n]], \quad (2.123)$$

where $E[\]$ is the expectation operator. The filter is adapted in function of the evolution of $\xi[n]$. This term is thus an indicator of convergence.

The desired-to-input cross-correlation vector $\mathbf{p}$ is defined in [9] as:

$$\mathbf{p} \equiv E[d[n] \cdot \mathbf{x}[n]] = [r_{dx}[0] \quad r_{dx}[1] \quad \dots \quad r_{dx}[L-1]]^T, \quad (2.124)$$

where the cross-correlation function between $d[n]$ and $x[n]$ is defined as:

$$r_{dx}[k] \equiv E[d[n] \cdot x[n-k]] \quad (2.125)$$

The auto-correlation function of $x[n]$ can be written in matrix form:

$$\mathbf{R} \equiv E[\mathbf{x}[n] \cdot \mathbf{x}[n]] = \begin{bmatrix} r_{xx}[0] & r_{xx}[1] & \cdots & r_{xx}[L-1] \\ r_{xx}[1] & r_{xx}[0] & \cdots & r_{xx}[L-2] \\ \vdots & \cdots & \ddots & \vdots \\ r_{xx}[L-1] & r_{xx}[L-2] & \cdots & r_{xx}[0] \end{bmatrix}, \quad (2.126)$$

where the auto-correlation function of $x[n]$ is defined as:

$$r_{xx}[k] \equiv E[x[n] \cdot x[n-k]] \quad (2.127)$$

Then, as shown in [9], it is possible to express $\xi[n]$ in function of $\mathbf{p}$ and $\mathbf{R}$:

$$\xi[n] = E[d^2[n]] - 2 \cdot \mathbf{p}^T \cdot \mathbf{w}[n] + \mathbf{w}[n]^T \cdot \mathbf{R} \cdot \mathbf{w}[n] \quad (2.128)$$

The performance function $\xi[n]$ is thus a quadratic function of the filter weights. Such a cost function is usually represented using only two weights in the filter: $L = 2$. It is possible to plot the Mean Square Error surface as a function of the two filter weights $w_0[n]$ and $w_1[n]$ as can be seen in Figure 2.11. $\mathbf{w}^o[n] = [w_0^o[n] \ w_1^o[n]]$ is the optimal coefficient vector at sample n that gives the minimum Mean Square Error $\xi_{\min}[n]$. Due to properties of quadratic functions, the optimal weight vector will always be unique, for any filter size L .

$\mathbf{w}^o[n]$ is an optimal set of control coefficients for which the MSE cost function is minimal. Then, vector differentiation of equation (2.128) gives a mathematical formulation of $\mathbf{w}^o[n]$:

$$\mathbf{R} \cdot \mathbf{w}^o = \mathbf{p} \quad (2.129)$$

Based on equation (2.129), the optimal filter can be found for any situation. The problem is that such a formulation requires to continuously compute the autocorrelation matrix $\mathbf{R}$ and cross-correlation vector $\mathbf{d}$, which represents an intense computation task. Moreover, this formulation implies that both the disturbance and error signals are perfectly stationary, which is rarely the case in practice. Recursive methods have been developed that enable the computation of the optimal filter in a progressive way requiring fewer computations. The Least Mean square algorithm is one method among many. It is introduced in the next section.

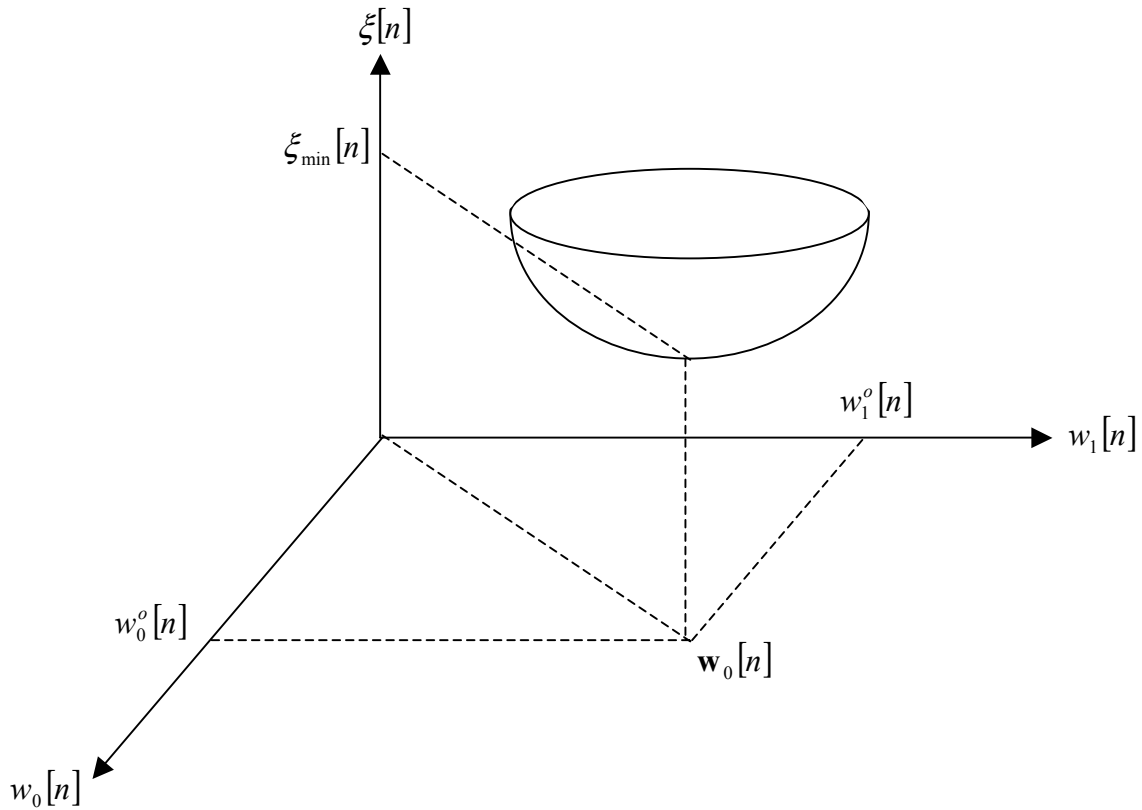


Figure 2.11 MSE surface. Simple case where $L=2$

b. LMS Algorithm Formulation

It was shown in the previous section that the MSE $\xi[n]$ is a quadratic function of the filter weights that can be seen as a positive concave hyper parabolic surface. To reach the minimum of such a function, gradient-based algorithms that will perform local estimates of the gradient and move incrementally toward the minimum of the surface can be used.

The method of steepest-descent constitutes a gradient-based algorithm. It is given by:

$$\mathbf{w}[n+1] = \mathbf{w}[n] - \frac{\mu}{2} \cdot \nabla \xi[n], \quad (2.130)$$

where $\nabla \xi[n]$ is defined as the vector of the MSE surface directional derivatives $\partial \xi[n] / \partial w_i$ and μ is a convergence factor that controls the stability and rate of descent to the minimum of the cost function. Based on equation (2.128), we can formulate this algorithm in function once again of $\mathbf{p}$ and $\mathbf{R}$:

$$\mathbf{w}[n+1] = \mathbf{w}[n] - \mu \cdot (\mathbf{p} - \mathbf{R} \cdot \mathbf{w}[n]) \quad (2.131)$$

But for the same reasons as before, such an algorithm is still difficult to implement since it requires many computations. Widrow [1970] simplified the expression of the mean square error $\xi[n]$ by estimating it using the instantaneous squared error $e^2[n]$:

$$\hat{\xi}[n] = e^2[n] \quad (2.132)$$

Then, the gradient of the Mean Square Error estimate, also called Least Mean Square, is given by:

$$\nabla \hat{\xi}[n] = 2 \cdot (\nabla e[n]) \cdot e[n] = -2 \cdot \mathbf{x}[n] \cdot e[n] \quad (2.133)$$

From equation (2.122), we deduce:

$$\nabla e[n] = -\mathbf{x}[n] \quad (2.134)$$

Then the gradient of the cost function estimate becomes:

$$\nabla \hat{\xi}[n] = -2 \cdot \mathbf{x}[n] \cdot e[n] \quad (2.135)$$

Substituting equation (2.133) into equation (2.130), we obtain the LMS algorithm also called stochastic gradient algorithm:

$$\mathbf{w}[n+1] = \mathbf{w}[n] + \mu \cdot \mathbf{x}[n] \cdot e[n] \quad (2.136)$$

Assuming a filter with L taps, the actual filter taps update is more simply expressed by:

$$w_l[n+1] = w_l[n] + \mu \cdot x[n-l] \cdot e[n] \quad l = \{0, 1, \dots, L-1\} \quad (2.137)$$

The convergence behavior of this algorithm is determined by the eigenvalues of the auto-correlation matrix of the input signal $x[n]$ and the size of the filter L . The maximum value for the convergence coefficient μ above which the algorithm may diverge can be approximated in function of the power of $x[n]$ and the number of filter coefficients (Elliot *et al*, 1989):

$$\mu_{\max} = \frac{1}{L \cdot E[x^2[n]]} \quad (2.138)$$

2.3.2. Practical Implementation of Adaptive Algorithms

2.3.2.1. Single Channel Feedforward Control

a. Introduction

We assume here that we want to identify an unknown plant whose dynamics are expressed in the z -domain by the transfer function $P[z]$:

$$P[z] = \frac{D[z]}{X[z]} \quad (2.139)$$

As we have seen before, the adaptive filter relates the controller output sequence $y[n]$ to the input signal $x[n]$:

$$W[z] = \frac{Y[z]}{X[z]} \quad (2.140)$$

The objective is to identify the plant dynamics by updating the adaptive filter $W[z]$ until it is similar to $P[z]$. The block diagram of such a system is shown in Figure 2.12.

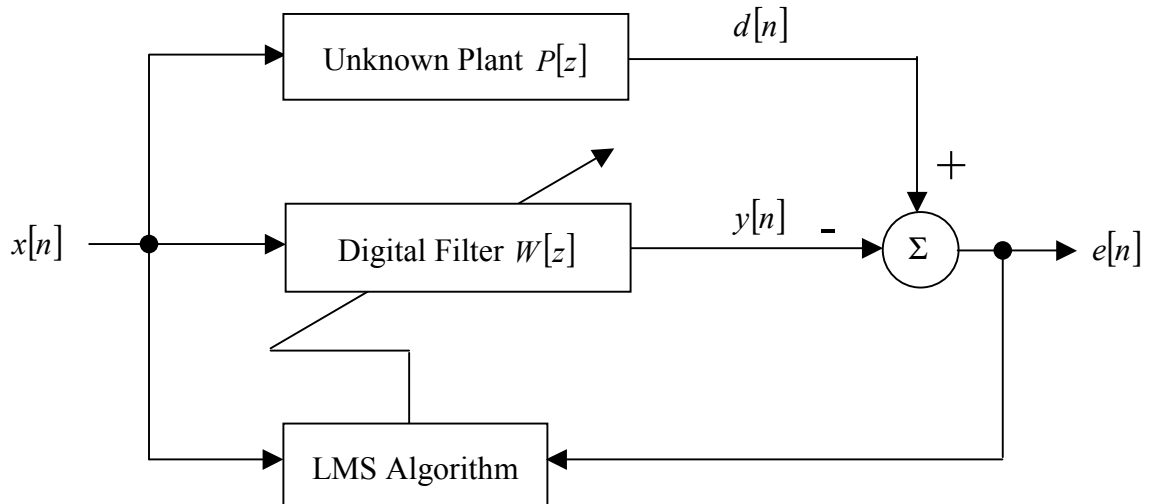


Figure 2.12 System identification block diagram

The adaptive filter $W[z]$ will converge toward $P[z]$ when we try to minimize the difference between the plant output sequence and the controller output sequence with the two systems being fed by the same input sequence.

b. Secondary Path Modeling and Effects

A simple application of single channel feedforward ANC is the control of noise propagating in ducts. We consider the problem of Figure 2.13.

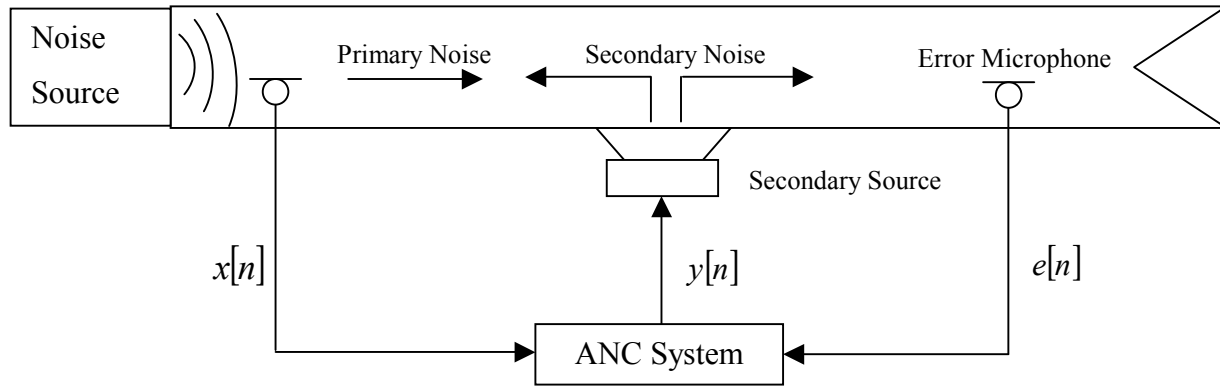


Figure 2.13 Single channel broadband feedforward ANC in duct

We consider a disturbance generated upstream and try to control it at the duct outlet using a speaker as secondary source. Microphones provide the error and reference signals. To ensure a good control, the reference signal should be well correlated with the disturbance so that the controller can properly construct a canceling waveform. The second most important objective is to ensure enough lag in the primary path (from the reference sensor position to the error) so that the controller has enough time to process the reference signal information and generate a propagating waveform that will interfere with the propagating disturbance such that noise cancellation is achieved at the error sensor location. The controller uses and generates discrete sequences of data. On the other hand, the goal here is to minimize acoustic quantities. Thus, there are number of transfer functions to take into account to accurately model the control system. These include the dynamics of the transducers used (microphones and speaker), the microphone signal conditioners, the speaker power amplifier, the D/A and A/D converters effects, the anti-aliasing and

reconstruction filters for the controller, and the acoustic path from loudspeaker to error microphone.

A block diagram corresponding to this complete control system is shown in Figure 2.14. We distinguish three different domains: acoustical, analog (electrical), and digital (electrical). The numerous elements constituting this block diagram can be combined together so that we just express the transfer functions (in the z-domain) relating the discrete reference, control and error signals. A simpler block diagram is shown in Figure 2.15.

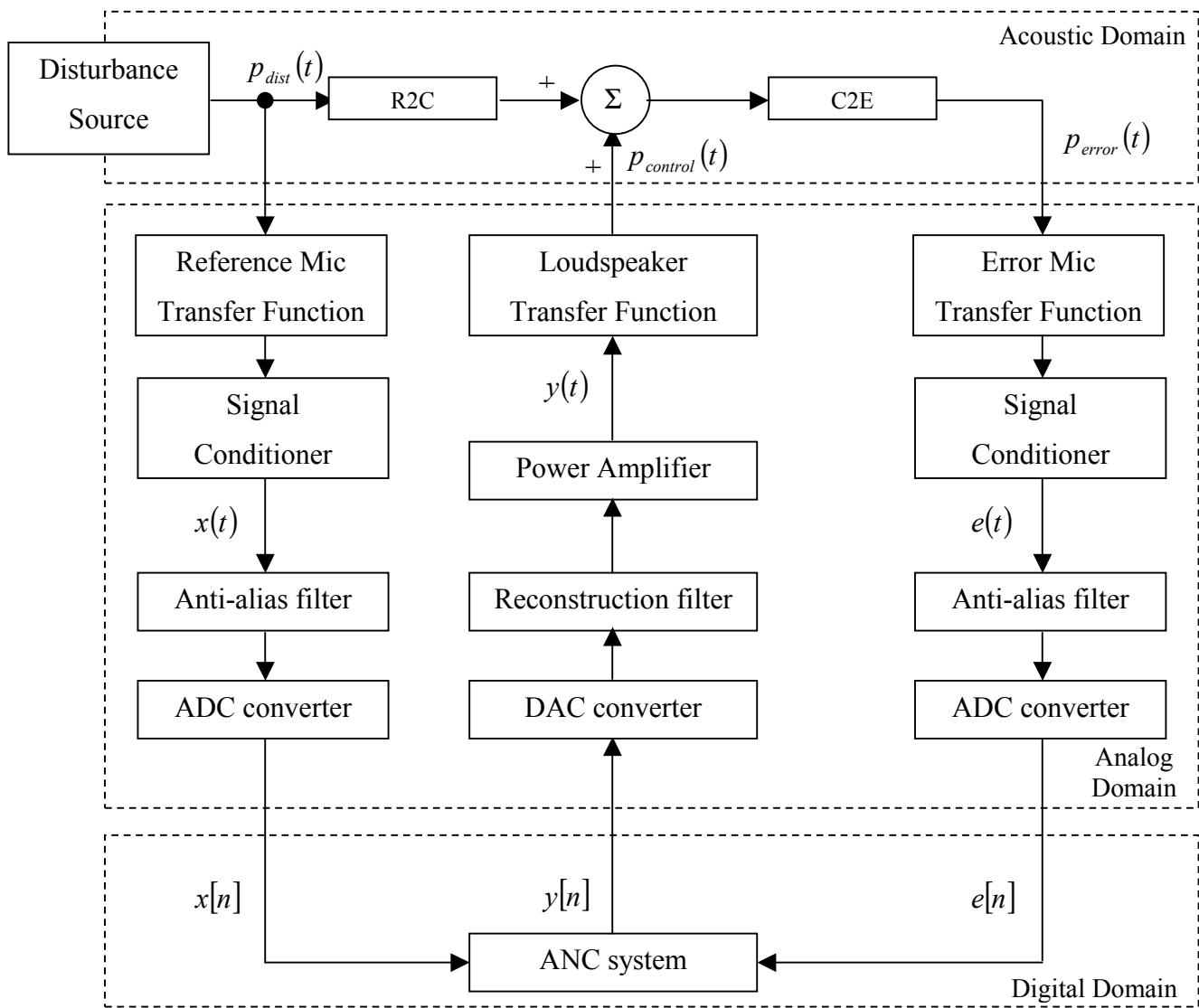


Figure 2.14 Complete block diagram of the single channel feedforward ANC in duct

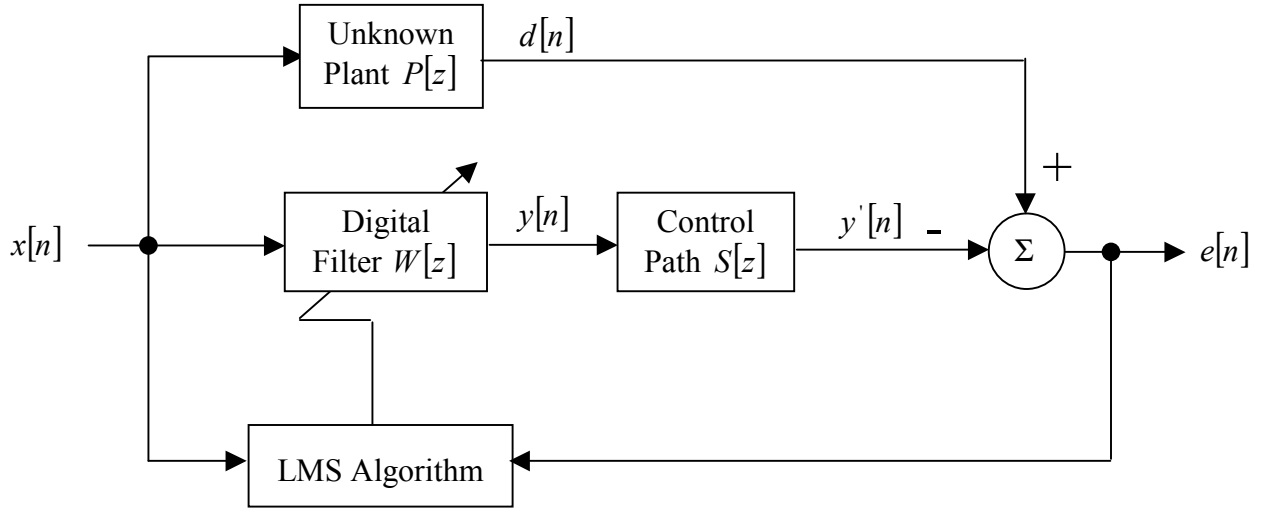


Figure 2.15 Secondary path modeling in single channel feedforward ANC

The z-transform of the error signal $e[n]$ can be expressed based on the block diagram of Figure 2.15:

$$E[z] = (P[z] - S[z] \cdot W[z]) \cdot X[z] \quad (2.141)$$

From equation (2.141), it is easy to deduce the optimal transfer function $W^o[z]$ that minimizes $E[z]$:

$$W^o[z] = \frac{P[z]}{S[z]} \quad (2.142)$$

Thus, the adaptive filter models the dynamics of the primary path $P[z]$ to generate a canceling noise at the error sensor and compensates for the effects of the secondary path $S[z]$. Nevertheless, when the filter adaptation is performed in the time domain using the LMS algorithm, the presence of a secondary path creates problems. Indeed, with the LMS algorithm (see equation (2.136)), it is assumed that the difference between the filter output and the desired response are directly available (compare Figure 2.10 and Figure 2.15), which is not the case in practice.

Regardless of the adaptation method (frequency or time domain), the secondary path can affect the controllability of a plant. In particular, at frequencies where the control path transfer function is zero, the optimal controller will tend to infinity (see equation (2.142)). This lack of control authority leads to an ineffective control. It is similarly observed that a zero in the primary path $P[z]$ will lead to an unobservable system at this particular frequency thus preventing any control.

One other important issue is the various time lags in the system. Referring to equation (2.142), we observe that the optimal control filter will be causal only when the lag induced by the secondary path is less than that in the primary path. This causality constraint is a limiting factor in broadband feedforward control systems. In case of pure tone control, where the disturbance presents highly predictable characteristics, causality is not an issue.

c. Filtered-X LMS Algorithm

It was shown in the previous section that the presence of a secondary path in the control system requires a modification of the LMS algorithm to update the adaptive filter. One of the solutions is to place an identical filter in the reference signal path [9]. This control architecture is known as Filtered-X LMS algorithm. A block diagram of such a control system is shown in Figure 2.16.

The Filtered-X LMS update algorithm is now given by:

$$\mathbf{w}[n+1] = \mathbf{w}[n] + \mu \cdot \mathbf{x}'[n] \cdot e[n], \quad (2.143)$$

where:

$$\mathbf{x}'[n] = \hat{s}[n] * \mathbf{x}[n] \quad (2.144)$$

The filtered reference signal $\mathbf{x}'[n]$ is obtained by convolution with an estimate $\hat{s}[n]$ of the secondary path impulse response $s[n]$.

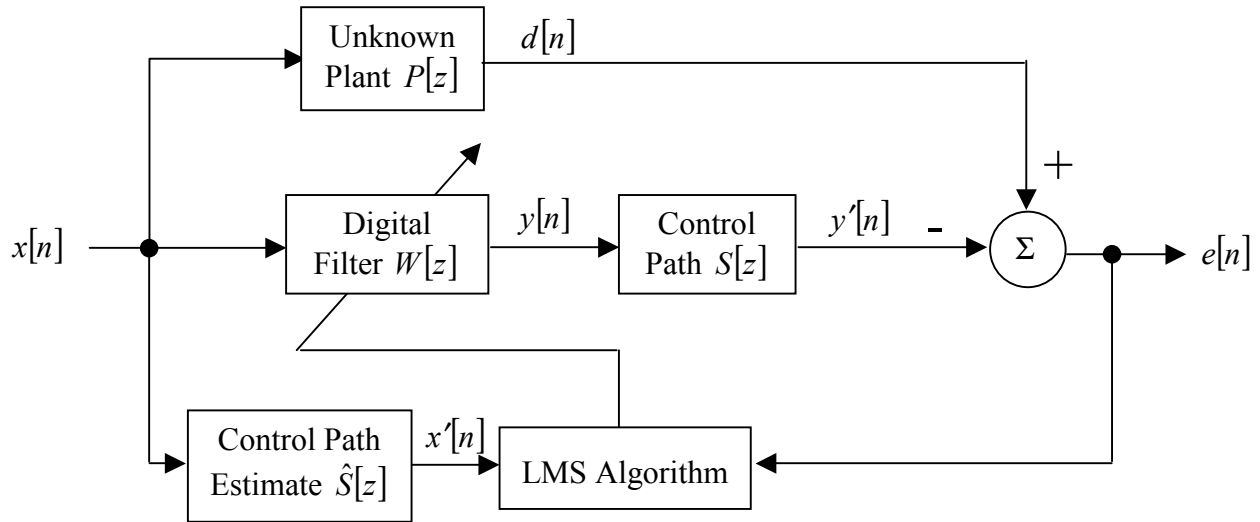


Figure 2.16 Block diagram of a single channel feedforward controller based on FXLMS

In practice, an identification phase precedes the control phase and permits the estimation of the secondary path. A block diagram of the operation is shown in Figure 2.17. The secondary source is fed with white noise, such that it generates a signal at the error microphone to be controlled. The optimal control filter is then an estimate of the control path that will be used during the control phase in FXLMS architectures.

The only limit of such a technique is that changes in the secondary path dynamics during the control phase will not be compensated for and thus might reduce the efficiency of the control. It has been shown [9] that errors in the secondary path identification affect the speed of convergence and optimal convergence factor. Nevertheless, for a SISO system, stability is still assured for phase errors in the range of ± 90 degrees.

In the case of Filtered-X LMS algorithm, the maximum value for the convergence coefficient μ is a function of the filtered reference signal power $E[x'^2[n]]$, adaptive filter size L , and overall delay in the secondary path Δ , in samples. The major limiting factor is usually the delay in the secondary path [24]:

$$\mu_{\max} = \frac{1}{L \cdot \Delta \cdot E[x'^2[n]]} \quad (2.145)$$

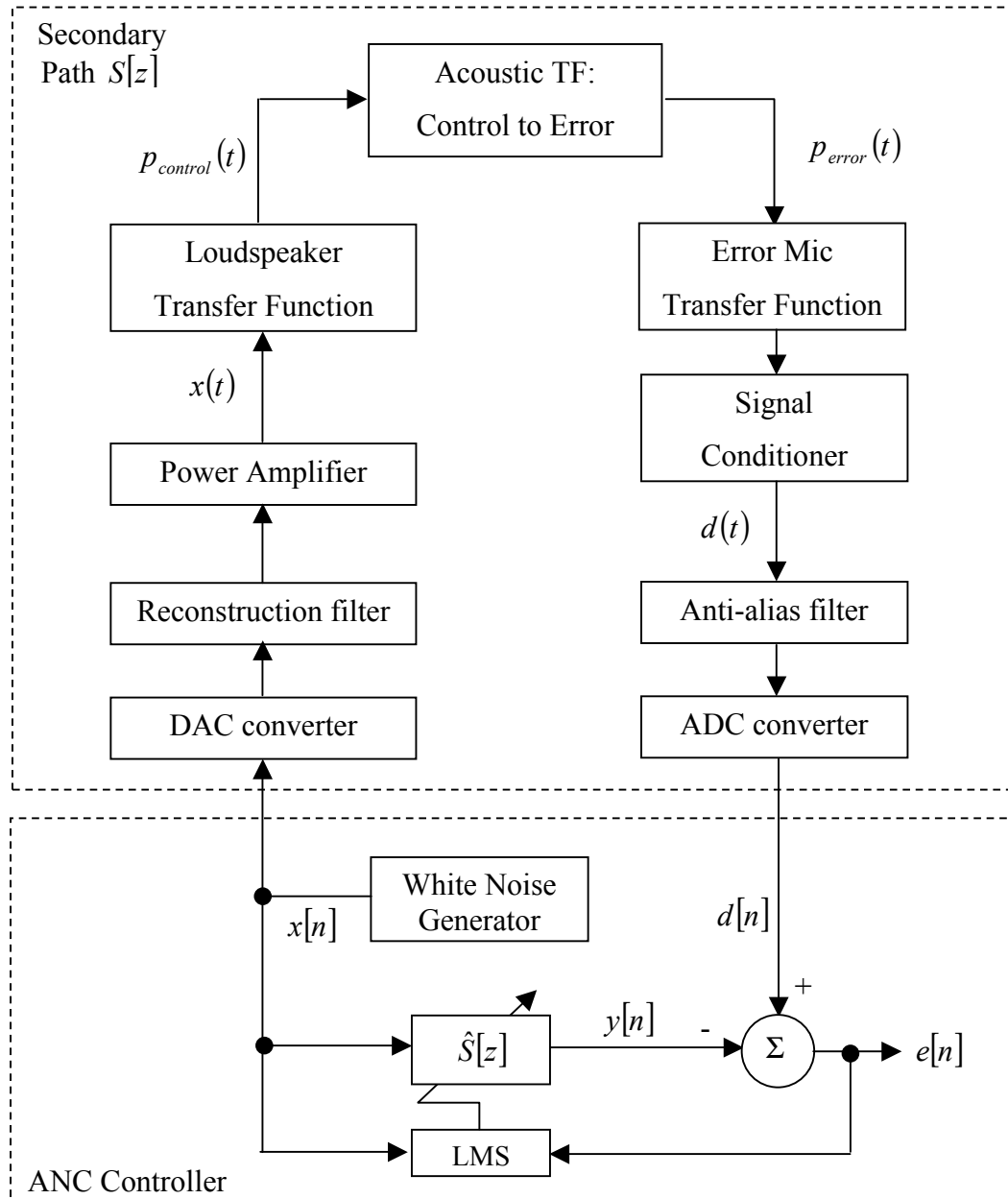


Figure 2.17 Secondary path identification block diagram

d. Measurement Noise Effects

In practice, measurement noise can be present in the reference and error signals. For example, in the case of control of fan noise, microphones in ducts can be subject to airflow-induced noise. We can quantify the effect of such problems by analyzing the block diagram of Figure 2.18.

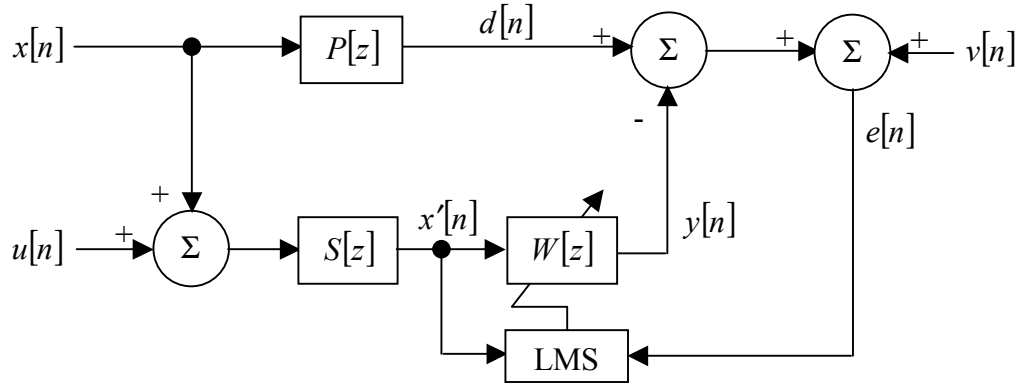


Figure 2.18 Block diagram of ANC system with noise on the input signals

For this situation, it can be proved [24] that the optimum filter is:

$$W^o[z] = \frac{S_{dx'}[z]}{S_{x'x'}[z]} = \frac{P[z]}{S[z]} \cdot \frac{1}{1 + \frac{1}{\rho[z]}}, \quad (2.146)$$

where $\rho[z] = \frac{S_{xx}[z]}{S_{uu}[z]}$ is the signal-to-noise spectral power ratio.

If the results of equation (2.146) are compared to the noise free case (equation (2.142)), we deduce that the performance of the controller will only be affected by the presence of noise on the input. The signal to noise spectral power ratio determines the limits of the control in a “noisy” situation. The controller will be performing a balance between canceling the disturbance noise and reducing the amplification of the measurement noise from the detection sensor.

e. Feedback Effects and Solutions.

The last problem we are dealing with in the case feedforward ANC is feedback effects from the control actuator to the reference sensor. For example, in the case of ANC in duct, the upstream reference microphone will most probably sense the noise generated by the control speaker located downstream as shown in Figure 2.13. One solution would be to use a reference sensor that is not sensitive to the control signal such as an accelerometer or tachometer. But, it will be shown that, in order to achieve a good broadband noise control (broadband noise in duct is due to air flow turbulences), a microphone is usually needed as reference signal.

A block diagram of a control system including such a feedback effect is shown in Figure 2.19.

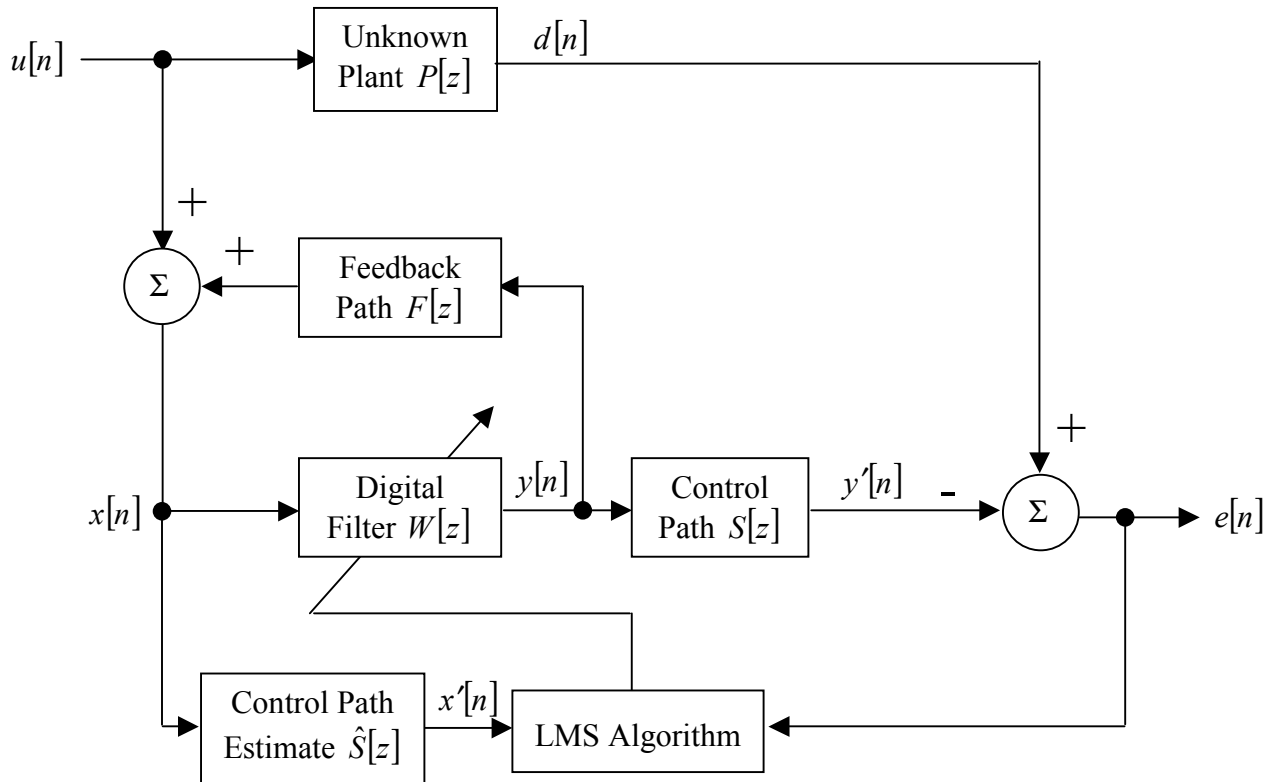


Figure 2.19 Block diagram of control system with feedback effect

Analyzing the previous diagram, the optimal filter can be deduced:

$$W^o[z] = \frac{P[z]}{S[z] + P[z] \cdot F[z]} \quad (2.147)$$

Then, the open-loop transfer function of the system with feedback becomes:

$$H_{OL}[z] = W[z] \cdot F[z] = \frac{P[z] \cdot F[z]}{S[z] + P[z] \cdot F[z]} \quad (2.148)$$

As explained in [9], the system will become unstable if the open loop transfer function accumulates a phase lag of 180 degrees, with an open loop gain above unity. Thus, a feedback loop in the system might prevent the controller from properly functioning.

One simple solution to get rid of the feedback effect is to include a feedback removal loop in the controller. During the identification phase, we will not only estimate the secondary path but also the feedback path based on the same procedure. In that case, two independent LMS controllers are used to estimate both paths as is illustrated in Figure 2.20.

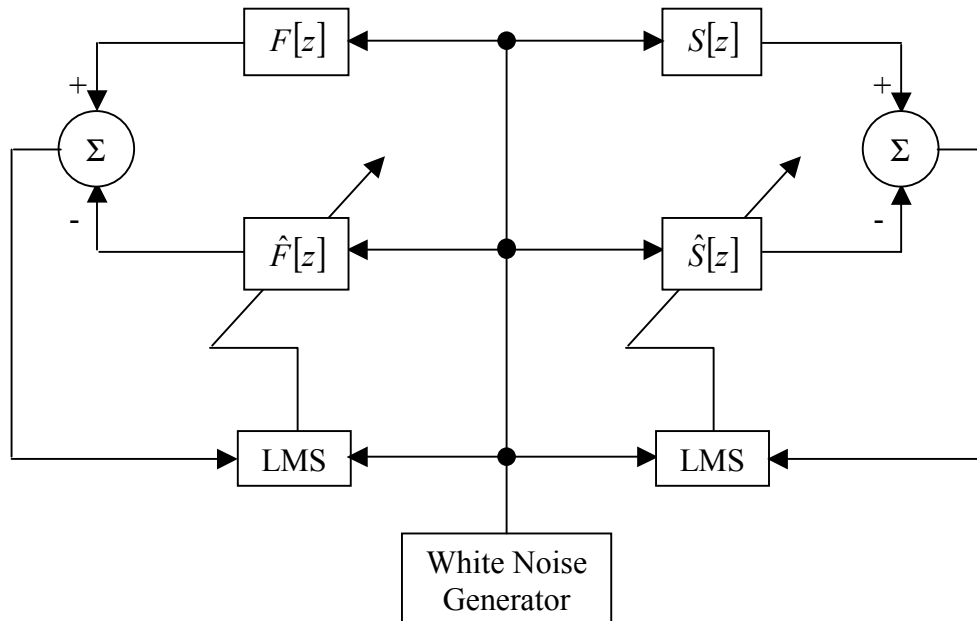


Figure 2.20 Simultaneous identification of the secondary and feedback paths

Finally, the complete single channel feedforward Filtered-X LMS controller system in duct is represented by the block diagram in Figure 2.21.

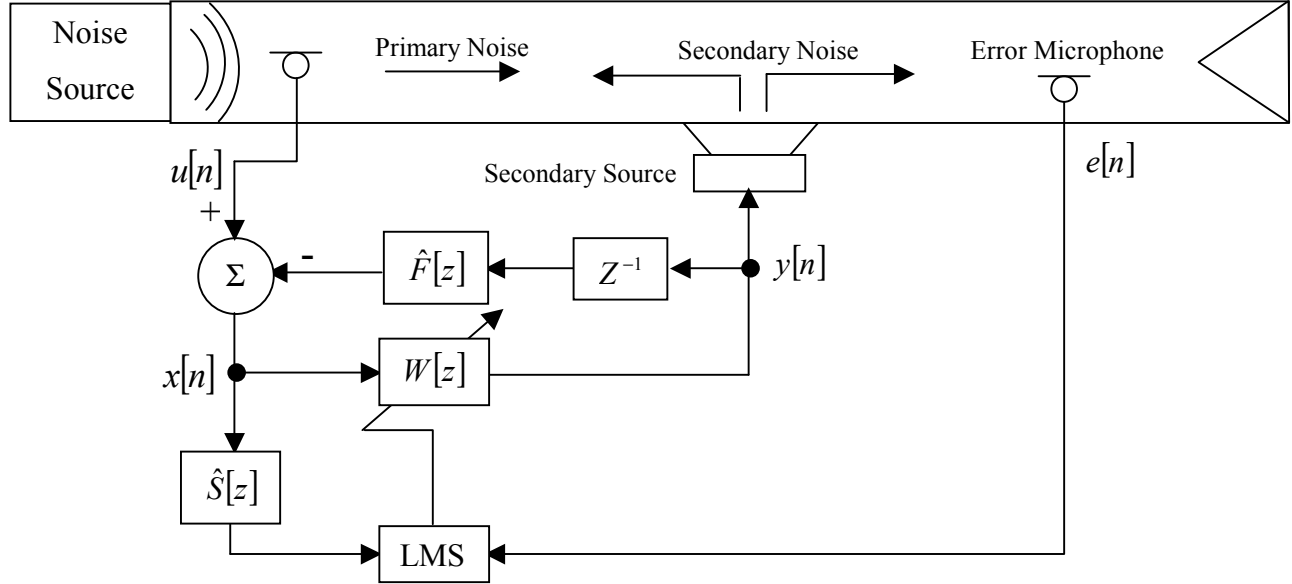


Figure 2.21 SISO feedforward ANC in duct. FXLMS with feedback removal filter

Up to this point, we have been interested in single input / single output controllers. This will indeed be the type of controller used for the simple Active Control of fan noise in ducts. But when in the case of the control of noise in a cavity, more than a single control actuator and error sensor will usually be needed. Moreover, a single reference sensor is very often not sufficient to ensure a proper control. Thus, in the next section, a generalized version of the Filtered-X LMS algorithm for multiple input / multiple output controllers will be introduced. Such a controller will be used for the local ANC of the PC noise at the user head location.

2.3.2.2. Multiple Channel Feedforward Control

a. Problem Formulation

It is assumed that the control system includes I reference sensors, J secondary sources and K error sensors. The feedback path is neglected since it will not be an issue in our MIMO control setups. The adaptive filters defined below will have L coefficients. The non-adaptive filters used to model the secondary path will be composed of M coefficients. The vector signals are manipulated in order to draw a simple block diagram as described in [9]:

$$\mathbf{x}[n] \equiv \{x_1[n] \quad x_2[n] \quad \cdots \quad x_I[n]\}^T \quad (2.149)$$

$$\mathbf{y}[n] \equiv \{y_1[n] \quad y_2[n] \quad \cdots \quad y_J[n]\}^T \quad (2.150)$$

$$\mathbf{d}[n] \equiv \{d_1[n] \quad d_2[n] \quad \cdots \quad d_K[n]\}^T \quad (2.151)$$

$$\mathbf{e}[n] \equiv \{e_1[n] \quad e_2[n] \quad \cdots \quad e_K[n]\}^T \quad (2.152)$$

The block diagram of the multiple channel control system is shown in Figure 2.22. $\mathbf{P}$ is the $(K \times I)$ primary path transfer function matrix. $\mathbf{W}$ is the $(J \times I)$ control filter matrix composed of $I \times J$ adaptive filters, each of length L . $\mathbf{S}$ is a model of the secondary path composed of $K \times J$ filters, each of length M .

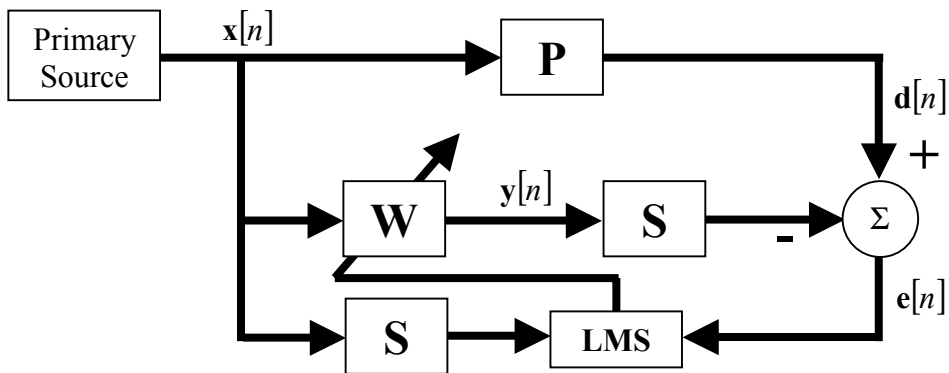


Figure 2.22 MIMO Filtered-X LMS control system

Each Transfer Function matrix defined earlier relates the input and output vectors. We take the example of the primary path matrix:

$$d_k[z] = P_{ki}[z] \cdot x_i[z] \quad (2.153)$$

Each sub-element of the matrices S and W is a discrete FIR filter defined as:

$$\mathbf{w}_{ji}[n] \equiv \{w_{ji0}[n] \quad w_{ji1}[n] \quad \cdots \quad w_{ji(L-1)}[n]\} \quad (2.154)$$

$$\mathbf{s}_{kj}[n] \equiv \{s_{kj0}[n] \quad s_{kj1}[n] \quad \cdots \quad s_{kj(M-1)}[n]\} \quad (2.155)$$

b. MIMO Filtered-X LMS Algorithm

We now express the output sequence of the adaptive filters that will drive the j^{th} secondary source. It is a linear sum of the I reference signals $x_i[n]$ optimally filtered by the set of filters $\{\mathbf{w}_{j1}[n] \quad \mathbf{w}_{j2}[n] \quad \cdots \quad \mathbf{w}_{jI}[n]\}$:

$$y_j[n] = \sum_{i=1}^I \sum_{l=0}^{L-1} w_{jil}[n] \cdot x_i[n-l] \quad (2.156)$$

Also, the k^{th} error sensor signal is composed of a linear contribution from the primary source and from each of the J secondary sources. The contribution of each control filter is performed via secondary paths that we model as FIR filters of length M:

$$e_k[n] = d_k[n] - \sum_{j=1}^J \sum_{m=0}^{M-1} s_{kjm} \cdot y_j[n-m] \quad (2.157)$$

By substituting equation (2.156) into equation (2.157), we obtain:

$$e_k[n] = d_k[n] - \sum_{j=1}^J \sum_{m=0}^{M-1} \sum_{i=1}^I \sum_{l=0}^{L-1} s_{kjm} \cdot w_{jil}[n-m] \cdot x_i[n-m-l] \quad (2.158)$$

Equation (2.158) can be simplified, provided the filtered reference signal $r_{kji}[n]$ is defined. It corresponds to the i^{th} reference signal filtered by $\mathbf{s}_{kj}$, the estimate of the secondary path between the j^{th} control source and the k^{th} error sensor:

$$e_k[n] = d_k[n] - \sum_{j=1}^J \sum_{i=1}^I \sum_{l=0}^{L-1} r_{kji}[n-l] \cdot w_{jil}[n], \quad (2.159)$$

where:

$$r_{kji}[n] = \sum_{m=0}^{M-1} s_{kjm} \cdot x_i[n-m] \quad (2.160)$$

The Mean Square Error (MSE) is defined in [9] as the expectation value of the sum of the squared error signals:

$$\xi[n] = E \left[\sum_{k=1}^K e_k^2[n] \right] \quad (2.161)$$

As we have seen before, a gradient-based algorithm will be used to minimize the cost function $\xi[n]$. In order to minimize the computational work when estimating the gradient of such a quadratic function of the filter weights, we now introduce an estimate of the cost function. Thus, we express the instantaneous cost function $\hat{\xi}[n]$:

$$\hat{\xi}[n] = \sum_{k=1}^K e_k^2[n] \quad (2.162)$$

The gradient of the cost function estimate defined in equation (2.133) is generalized here for the multiple errors case. For each control filter weight, the derivative of the cost function is given by:

$$\frac{\partial \hat{\xi}[n]}{\partial w_{jil}[n]} = 2 \cdot \sum_{k=1}^K \left(\frac{\partial e_k[n]}{\partial w_{jil}[n]} \right) \cdot e_k[n] \quad (2.163)$$

From equation (2.159), we deduce that:

$$\frac{\partial e_k[n]}{\partial w_{jil}[n]} = -r_{kji}[n-l] \quad (2.164)$$

Then, equation (2.163) becomes:

$$\frac{\partial \hat{\xi}[n]}{\partial w_{jil}[n]} = -2 \cdot \sum_{k=1}^K r_{kji}[n-l] \cdot e_k[n] \quad (2.165)$$

Inserting the previous result in the steepest descent algorithm, first introduced in equation (2.130) for the single error case, we obtain the multiple Error Filtered-X LMS algorithm defined for the first time by Elliott and Nelson in 1985 [9]:

$$w_{jil}[n+1] = w_{jil}[n] + \mu \cdot \sum_{k=1}^K r_{kji}[n-l] \cdot e_k[n] \quad (2.166)$$

This algorithm will be used for the local control of PC noise since multiple reference sensors (both structural and acoustical), control speakers and error microphones to will be required to maximize the achievable control as well as the size of the zone of quiet surrounding the error sensors.

c. Exact Least Squares Solutions

We consider the exact Mean Square Error defined in equation (2.161). It was shown that this cost function had a unique minimum leading to a single optimal filter in the case of a single input / single output controller.

We now consider a multiple channel controller with J secondary sources and K error sensors. The exact least square solution we are seeking here is still the minimum of the error surface. By expressing equation (2.157) in the z-domain, the problem can be viewed as a set of K error equations to solve were the unknowns are the J secondary signals $y_j[z]$:

$$e_k[z] = d_k[z] + \sum_{j=1}^J S_{kj}[z] \cdot y_j[z] \quad (2.167)$$

If the number of error sensors K is greater than the number of secondary sources J , the problem is over-determined since there are more equations to solve than unknowns. On the other hand, using more secondary sources than error sensors lead to an undetermined problem, which can pose problems in practice since we can't theoretically find the optimal solution. The ideal case where $K=J$ can lead to the same troubles if the 2 sources or 2 errors are close together.

In [9], Nelson and Elliott relate this problem to the conditioning of the secondary path matrix $\mathbf{S}$. Indeed, they express the cost function ξ in the Hermitian Quadratic form:

$$\xi = \mathbf{e}^H \mathbf{e} = \mathbf{d}^H \mathbf{d} + \mathbf{d}^H \mathbf{S} \mathbf{y} + \mathbf{y}^H \mathbf{S}^H \mathbf{d} + \mathbf{y}^H \mathbf{S}^H \mathbf{S} \mathbf{y}, \quad (2.168)$$

where H is the Hermitian operator (complex conjugate of the transpose matrix). If we look for the optimal set of inputs $\mathbf{y}_o$ to the secondary sources, which minimize the cost function, we obtain the solutions from Table 2.1:

Table 2.1 Exact solution to the MSE minimum in MIMO configuration

Case:	Optimal set of control signals:
Over-determined: $K > J$	$\mathbf{y}_o = -(\mathbf{S}^H \mathbf{S})^{-1} \mathbf{S}^H \mathbf{d}$
Fully determined: $K = J$	$\mathbf{y}_o = -\mathbf{S}^{-1} \mathbf{d}$
Underdetermined: $K < J$	$\mathbf{y}_o = -\mathbf{S}^H (\mathbf{S} \mathbf{S}^H)^{-1} \mathbf{d}$

For each case, the computation of the optimal solution requires the inversion of a matrix. If this matrix is ill conditioned (determinant nearly equal to zero), it will not be invertible. In such a case, it is impossible to find the optimal solution.

For the over-determined case, the matrix $\mathbf{C}^H \mathbf{C}$ is ill conditioned when two of its columns are nearly equal which physically corresponds to the case where two secondary sources are close together. For the fully determined case, the matrix $\mathbf{C}$ is ill conditioned when either two sources or two errors are close together (two columns or two rows are nearly equal). In the underdetermined case, the matrix $\mathbf{C} \mathbf{C}^H$ is ill-conditioned when two error sensors are close together (two columns are nearly equal).

Chapter 3. Noise Sources Identification

The objective of this chapter is to bring some understanding regarding the noise radiated from modern personal computers. Before making any decision regarding what type of noise control treatments to be applied and on which item, the noise sources in the PC need to be identified, characterized, and ranked. We will see that Active Noise Control is a technique that is most efficient at low frequencies (below 1000Hz) and can be best implemented when the noise sources are reduced to simple acoustic sources. Thus, the acoustical properties of the identified noise sources need to be studied in detail in order to determine the suitability of active techniques and choose the appropriate control scheme. Finally, it is most likely that the applied control treatment will combine passive and active technology. Passive treatment could be acoustic or structural. On the other hand, our work on active techniques will be limited to the active control of the noise radiated by the PC. We will find out that both airborne and structure-borne noise radiation paths are present in the PC. Thus, in order to select the appropriate noise control treatment for each problem, multiple noise radiation paths must be investigated.

In the first part of this chapter, the Personal Computer used for the past two years experiments will be presented. Some of the acoustical properties of the various noise sources will be determined based on sound power level as well as directivity measurements. Since the casing might significantly affect the noise radiated by each noise source and even constitute new radiation paths, their interaction will be investigated. An intensity survey performed on the casing with all the noise sources operating will help to characterize the overall noise radiation and detect any radiation hot spot that is to be controlled. Finally, the noise sources will be ranked in order to determine which sources are to be targeted in priority. An important goal in this part of the work will be to discern the major radiation paths (airborne or structure-borne) such that efficient noise and vibration control treatments can be applied.

3.1. Dell Precision Workstation 410 Computer

3.1.1. Geometry

The computer that forms the being of this study is the Dell Workstation 410. The standard ATX casing of this unit is classified as a mini-tower. The overall dimensions of the machine are 44 by 22 by 44 cm in height, width, and depth respectively, and the production weight is 13kg. As can be seen in Figure 3.1, two of the casing covers are removable by the means of release tabs. The front bezel is made of plastic where as the side panels are composed of steel sheet metal assembled against the plastic cover for electromagnetic isolation purposes. The modified ATX chassis is also made of steel sheet metal, and machined so that various PC hardware such as the fans, power supply block, and disk drives can be easily accessed and replaced.

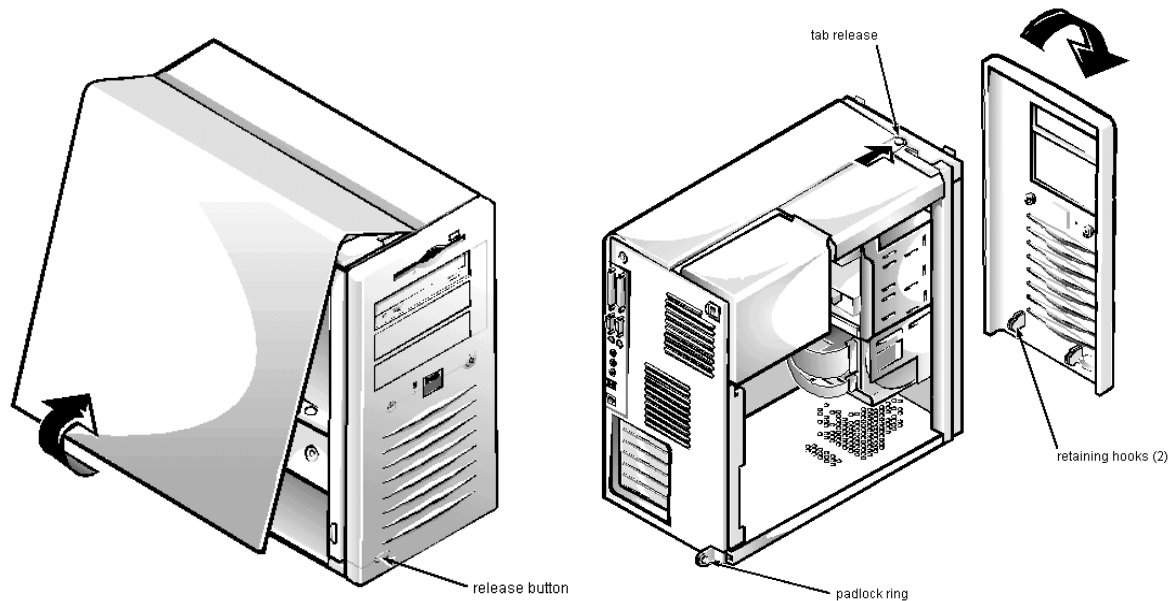


Figure 3.1 Overall view of the Dell Precision Workstation 410 casing

3.1.2. Hardware Components

The main computer components are the power supply block, the system board and the peripherals such as floppy, CD-ROM, and hard drive. The floppy and CD-ROM drives are mounted on the external drive bays as shown in Figure 3.2. The single hard-drive incorporated in the machine is mounted in the internal drive bays. The system board, which houses all the external hardware connectors as well as the single processor unit, is vertically mounted on the left part of the PC as shown in Figure 3.2. The power supply block, which includes a cooling fan, constitutes a removable unit mounted on the top of the PC, at the rear. We can see apertures on the rear panel of the chassis, next to the power supply block and below it. These are the air outlets of the fans. Indeed, as will be shown in the next section, a chassis mounted fan is dedicated to the processor cooling as well as venting of the PC casing.

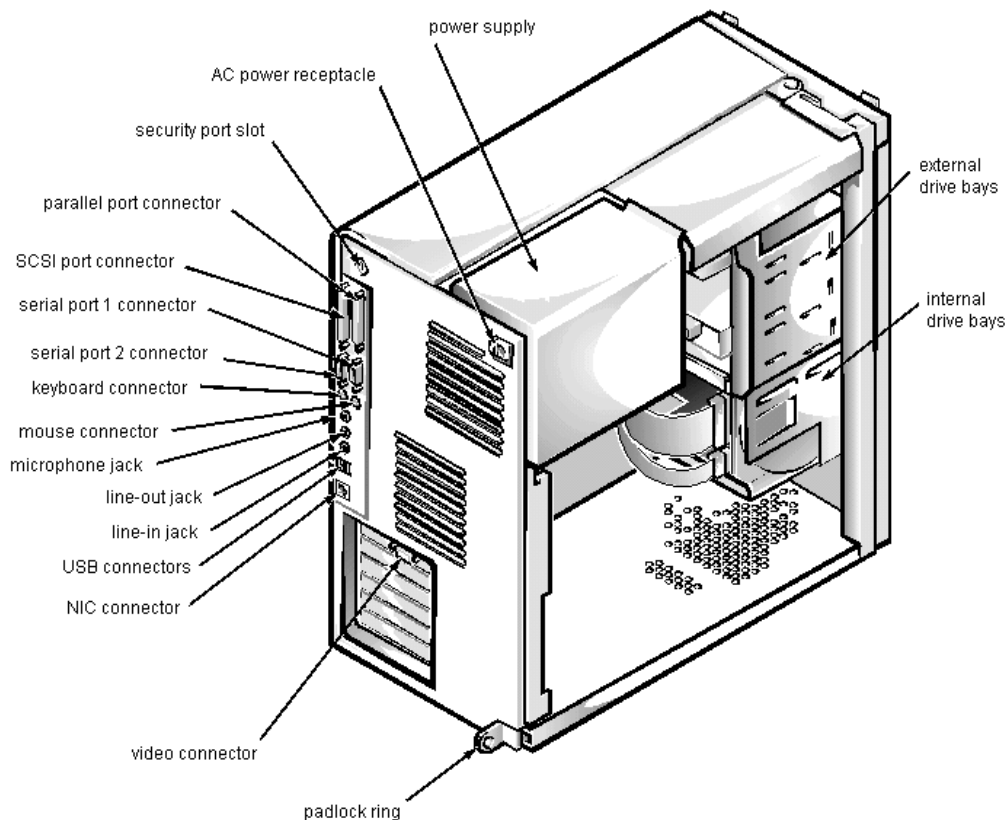


Figure 3.2 Hardware configuration in the Workstation 410

3.1.3. Identified Noise Sources

In this project, we are exclusively interested in the stationary noise sources of the PC. For example, we take the example of the hard drive. This unit is composed of several discs that are constantly spinning at a nominal speed as soon as the drive is powered up. But, when the system is retrieving information stored on such a device, the seeking process will lead to the generation of additional transient type noise. For practical reasons in Active Noise Control, such transient type noise is not controlled because its non-stationary characteristics can usually not be tracked by a typical adaptive controller whose adaptation rate is best suited for slow time-varying processes. The external drives such as floppy and CD-ROM drives do not spin in idle mode, thus they do not constitute stationary noise sources either. Thus, those sources are not target for an adaptive control system.

Consequently, the noise sources determined *à priori* and identified in the following tests are the power supply block with its integrated fan, the processor fan, and the hard drive. All these components run at a nominal speed as soon as the motherboard is powered up. Both fans are not temperature controlled, and their speed remains constant regardless of the PC operating conditions. The operating system used does not permit to power down some devices when not used (such as the hard drive).

3.2. Sound Power Level Measurements

There are two widely used methods for the determination of a source sound power level requiring the use of either an anechoic or a reverberant chamber. Both methods were used for the following measurements. Nevertheless, when applying the reverberant chamber method, corrections need to be applied in order to take into account the absorption of the room (c.f. references [16], [17] and [26]). In particular, the reverberation time of the room needs to be measured so that the radiated sound power level can then be inferred from the measurements. At the time of testing, the reverberation time of the room was not known with sufficient confidence to accurately estimate the sound power levels. Thus, the integrated sound power level results listed below are exclusively based on anechoic

chamber measurements. For further information regarding the testing procedures, one should refer to section 2.2.1.

3.2.1. Experimental Setup.

3.2.1.1. Power Level Measurements Setup.

For the measurements in the anechoic chamber, the computer was placed over a reflecting surface. A hemispherical array of radius 1 meter was used to dispose microphones at strategic locations as shown in Figure 3.3. These microphones (a total of ten) are specially located so that they represent an equal surface of the hemisphere surrounding the PC. The pressure amplitude squared (proportional to the intensity level, assuming we are in the far field of the radiation: $kd \gg 1$ with k the wavenumber and d the distance to the source) is summed over the hemisphere surface in order to deduce the source power level.

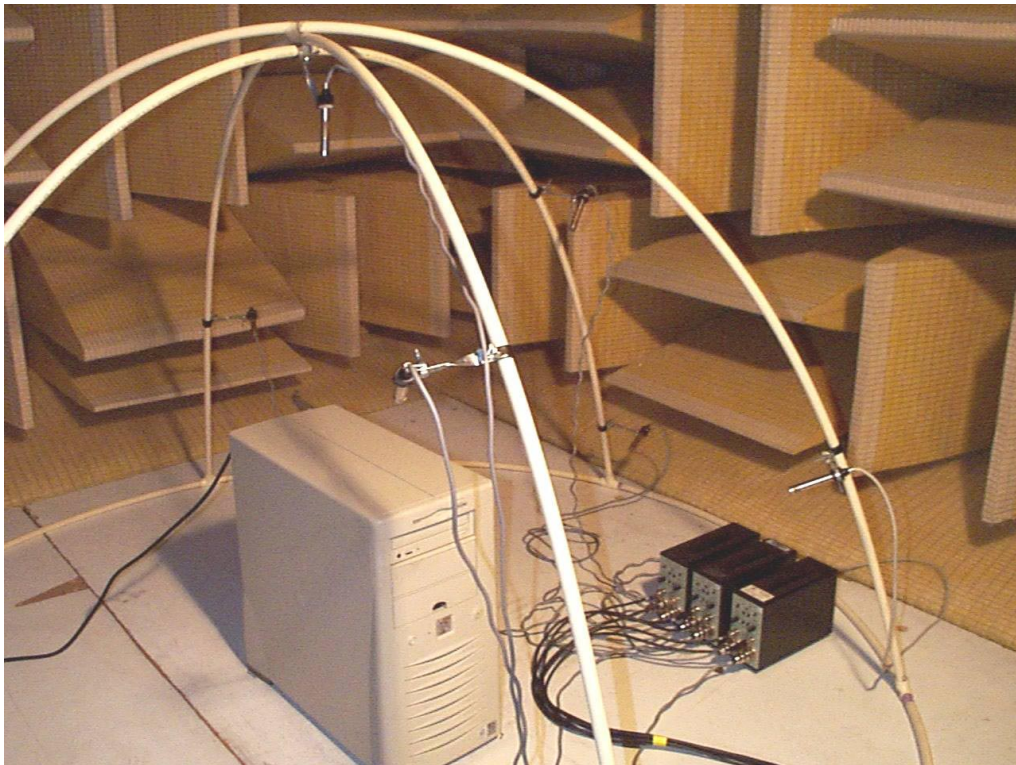


Figure 3.3 Sound power level measurement in one of the VAL anechoic chambers

For the measurements performed in reverberant chamber, the total energy in the room increases until reaching equilibrium. At the equilibrium, the amount of power dissipated by the room equals the amount of power radiated by the source. By measuring the steady state sound pressure level in these conditions, the sound power level of the source can then be determined. In theory, the sound pressure is constant throughout the room, since the sound field is assumed diffuse. In practice this is rarely true and the sound pressure level is measured at different locations (at least eight) in order to spatially average the results obtained. Figure 3.4 shows the Dell computer in a VAL reverberation chamber.

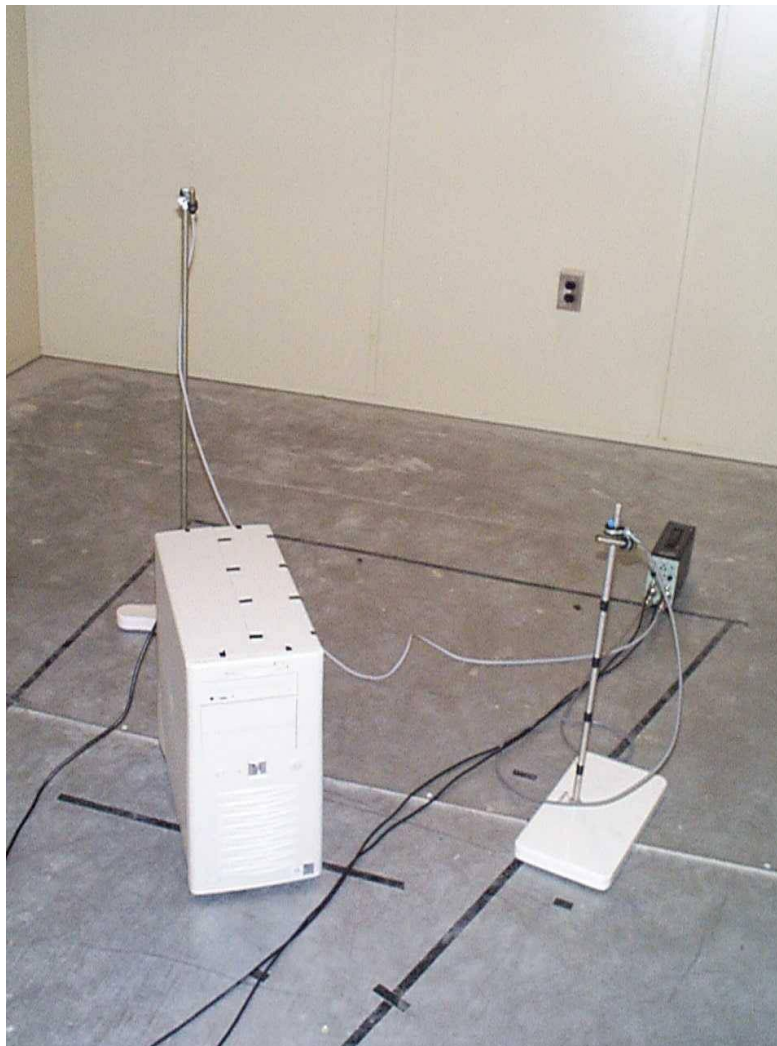


Figure 3.4 Sound power level measurement in a VAL reverberant chamber

3.2.1.2. Practical Measurement Problems

The characterization of the computer noise sources was difficult. One of the problems was the overall low sound power radiated by the computer. Moreover, results were difficult to replicate because the source noise characteristics would change in time. Those two major issues are detailed below.

a. Low Level Signals

The identified noise sources generate low sound pressure levels, of the order of 40 to 45dBA. Most microphones will generate low voltage output at such low-pressure levels and the signal to noise ratio will likely be poor. In order to limit quantization noise during the analog to digital signal conversion process, high voltage gains need to be applied to the microphone signal. Those voltage amplifiers can degrade the signal to noise ratio of the acquisition chain.

Two types of microphones were used for this project. A few B&K ½” microphones (Type 4166) were available. This high precision equipment presents a large signal to noise ratio but requires the use of costly preamplifiers and power supplies. Thus, less precise equipment such as PCB/Acousticel microphones (type TMS 130A/B), was used due to its availability in large quantities (as needed by the hemispherical array for example).

These PCB microphones have the inconvenient of presenting a lower signal to noise ratio than the B&K units, which is a problem for low level signal measurements such as PC testing. When PCB units are used, the measurements can be significantly corrupted by extraneous noise. The PCB microphones are also more sensitive to 60Hz noise than the B&K sensors. Both measurement techniques (in the anechoic chamber or the reverberant chamber) require the use of multiple microphones. The trade off was then to either use a small number of B&K microphones with high SNR and perform the data acquisition in several sets or use a large amount of PCB/Acousticel microphones and record all the

signals simultaneously. It will be shown in the next section that recording the data over multiple time frames presents some problems due to the limited stationarity characteristics of the noise sources measured.

b. Stationarity and Reproducibility Issue

During the phase of identification of the noise sources, we started our measurements using high precision B&K microphones. Since only a small quantity of those sensors were available at a time, the sound power measurements based on multiple microphone locations could not be performed at once. Theoretically, the power level should be computed on signals all acquired at the same time except if the source under test is perfectly stationary. We found out that the noise sources identified were, in reality, changing characteristics during the data-acquisition period. For example, the fans BPF and harmonics would shift in frequency depending on the data acquisition set. For this reason, the B&K units were soon replaced by the PCB/Acousticel microphones, that would enable us to record all the microphones at once. We could then ensure homogeneous results among the set of data, but at the expense of a significant data corruption due to the limited signal to noise ratio of the PCB units.

This raises one other issue, which is the reproducibility of the measurements. In practice, we performed multiple power level measurements, in different chambers (anechoic as well as reverberant), using several types of microphones (B&K or PCB/Acousticel). The power levels measured would vary from 1 to 3dB, depending upon the test room and the source characteristics at the time of the measurements. The broadband characteristics of the noise sources are usually reproducible. On the other hand, the harmonic content for the fans noise particularly, changes regularly in frequency and amplitude. Those variations originate from changes in blades rotational speed, which could be due to either instability in the DC voltage supply or changes in the airflow resistance. As we will observe, the way the noise sources are mounted in the PC casing does greatly affect the measurement. Thus, manipulating the sources (removing it from the PC casing and putting it back in afterwards) between two acquisitions could lead to different results.

One other explanation for the variations observed regarding the pure tones concerns the anechoic chamber measurements. Ideally, the intensity level should be integrated over the whole surface surrounding the noise source, not only at few locations. When using only 10 microphones, we implicitly assume that the noise source is very omni-directional, which may not be true for some of the pure tones radiated.

3.2.1.3. Data Acquisition and Processing

Since the sound levels measured are significantly low, the microphone signals were amplified with Ithaco analog filters (gain of 100). These units were also used as anti-aliasing filters. Aliasing occurs when the signal to be converted contains information at frequencies above half the sample rate. Thus the second order low pass filter had a cut-off frequency set at a third of the data acquisition sample rate. The signals coming from the various microphones were always acquired in the time domain at a sample rate of usually at least 15kHz, during a period of 10 seconds, using a 16-channel National Instruments data acquisition card. A Windows interface controlling the acquisition settings has been developed in the VAL Labs using the Labview programming tool. The number of channels, acquisition sample rate and acquisition duration can be selected. Analog anti-aliasing filters built in the acquisition system can also be set to provide supplementary voltage gain and anti-alias filtering in case we do not use external low-pass filters.

The acquired time signals are A-weighted [26] using digital filters programmed in Matlab. As can be seen in Figure 3.5, the A-filter corresponds to a band-pass filter weighting both ends of the frequency range of interest to compensate for the human ear sensitivity since the perceived loudness of a signal is frequency dependent [26].

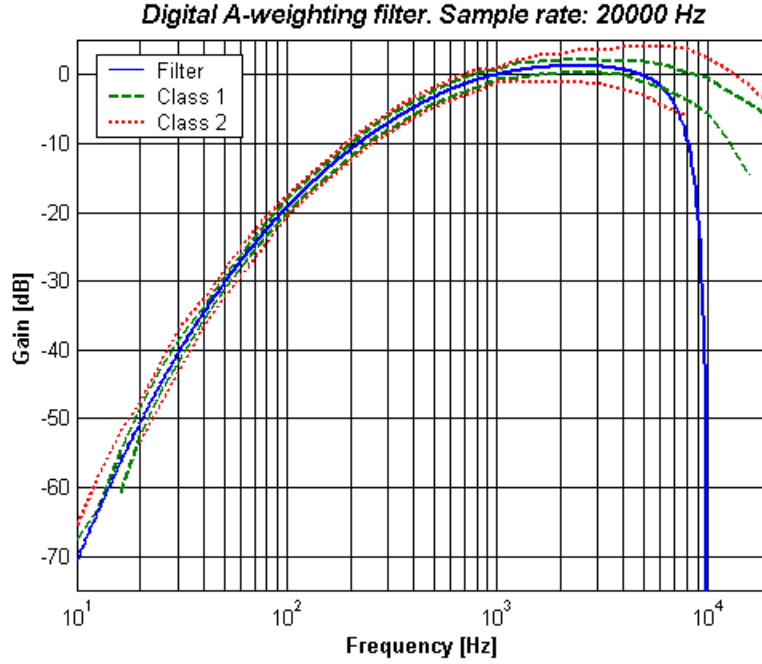


Figure 3.5 A-weighting filter gain versus frequency curve

The power spectra computed below are based on the algorithms presented in section 2.1.1. The FFT's are performed on blocks of the original data so that averages can be made to estimate the power spectrum. The number of averages determines the blocks size, which is also related to the frequency resolution of the result. Typically the data blocks will have to be truncated such that the length is a power of two in order to minimize the computation time when using the Fast Fourier Transform algorithm. The frequency resolution is given by:

$$\Delta f = \frac{1}{T_t} \cdot nd = \frac{F_s}{N_t} \cdot nd ,$$

where:

nd is the number of averages

T_t is the truncated acquisition period (seconds)

F_s is the acquisition sample rate (Hz)

N_t is the truncated block size (samples)

3.2.2. Results

3.2.2.1. Power Supply Block Fan

a. Location

The removable power supply block is shown in Figure 3.6. Since it integrates a fan spinning continuously when the computer is powered up, the power supply constitutes a source of noise of major interest. The fan has a diameter of 80 mm and its powered by 12V DC voltage. It incorporates seven blades while the effective blowing area is approximately 45 cm².

The fan blows air out of the PC casing through the power supply block. The “cool” air, coming from inside the PC, and enters the power supply block from the opposite side to the fan. Thus, there is a flow circulation from one end of the power supply block to the other. We see in Figure 3.6 that the hot air is then expelled out of the PC casing through the rear panel. The noise generated by this particular fan can now be characterized based on the following power level measurements.

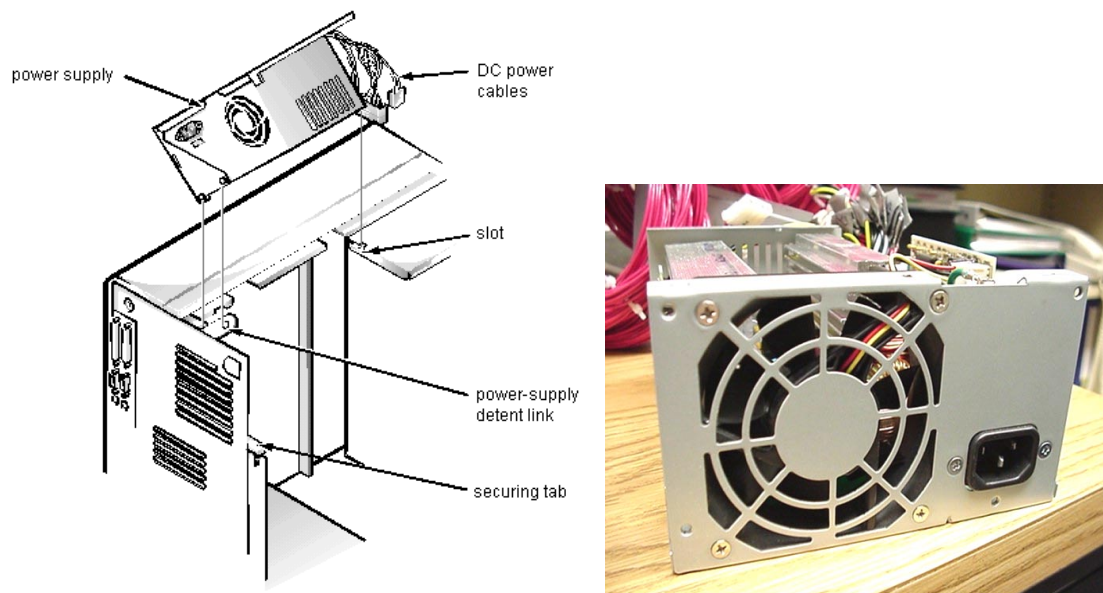


Figure 3.6 Power supply block with integrated fan

The sound characteristics of the power supply block have been measured when mounted in the casing as well as out of the casing. Some of the results shown below correspond to measurements in a VAL anechoic chamber. The computer was placed on a large rigid panel and surrounded by the test hemisphere with ten PCB microphones. These acoustic sensors have a reasonable sensitivity (approximately 150 mV/Pa). Unfortunately, their signal to noise ratio is limited since they are commonly used for loud noise sources characterization. Moreover, these particular PCB sensors are sensitive to 60Hz noise. In the following measurements, the background noise of the chamber is included. The acoustical background noise measured in the anechoic chamber is actually overestimated due to the presence of measurement noise in the recorded signals.

Some other results presented below correspond to measurements in one of the VAL reverberant chambers, where B&K microphones (Type 4166) were used. As will be seen in the next plots, signal to noise ratio was not an issue when using such acoustic sensors.

b. Sound Characteristics in PC Casing

The power supply block noise is first characterized while mounted in the PC casing. In order to only measure the power supply block fan related noise, the other major noise sources, i.e. the processor fan and the hard drive, have to be powered down. In Figure 3.7, the repeatability issue is illustrated by plotting the results of measurements of the power supply block fan noise, taken at two different dates, using the same equipment, with an identical noise source configuration. As can be seen, the overall power level slightly varies with high level tones changing in amplitude with time. The broadband characteristics remain homogeneous in terms of frequency distribution. The blue curve represents the most common type of power spectrum obtained for the power supply block fan, i.e. a broadband noise important up to 600Hz and some pure tones related to the Blade Passing Frequency (BPF) and harmonics. The BPF is usually 10 to 15dB above the broadband level, and up to 7dB for the harmonics.

Overall, the power supply block, when mounted in the PC casing, radiates a sound power level between 42 and 44dBA, with most of the energy contained in low frequency tones and broadband noise below 600Hz. The radiation is negligible above a 1000Hz as will be observed in the next section, which means that such a noise source is suitable to be controlled using active devices. The pure tones are related to periodic nature of the rotation of the fan blades, and the broadband noise is most likely due to turbulences in the airflow (the geometry is fairly complex as we have seen before).

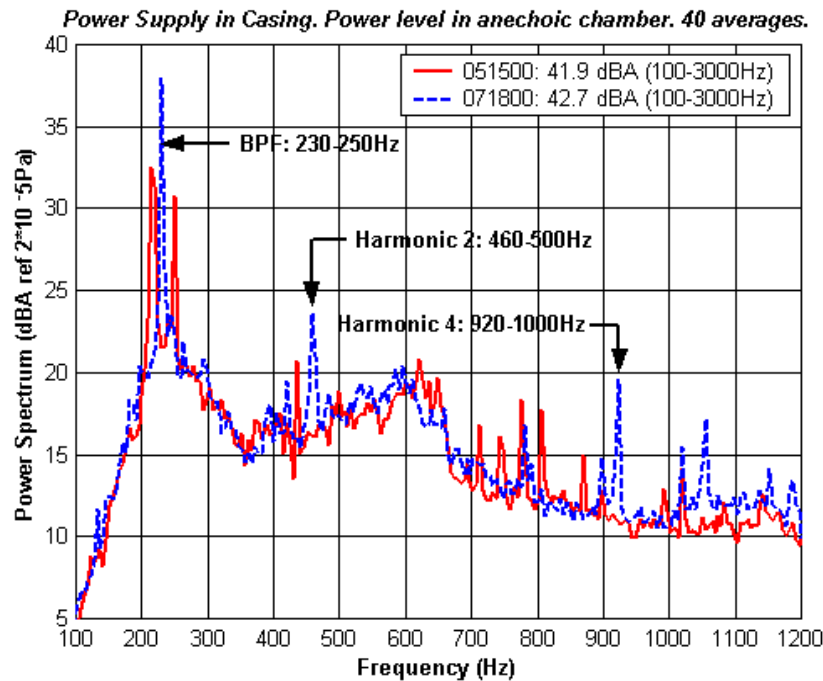


Figure 3.7 Power supply block sound power level. Repeatability issue.

c. PC Casing Effect.

It is important to investigate the influence of the PC casing on the sound power level radiated by the power supply block fan. The results for measurements made in the anechoic chamber using PCB microphones are shown in Figure 3.8 where as those in the reverberant chamber using B&K microphones are shown in Figure 3.9. It can be observed that the background noise is lower for the measurements in reverberant chamber, which is due to the higher signal to noise ratio of the B&K sensors.

The second remark concerns the time varying characteristics of the measured noise source. The measurements in reverberant and anechoic chambers were not taken the same day. Slight differences in the results from one measurement to the other are observed.

Still, the measured sound power level of the power supply block in and out of casing can be compared. In both cases, most of the power is radiated at frequencies below 1000Hz. The broadband characteristics of the power spectra vary in strong relation with the test configuration. This may be due to the change in airflow constraints. Regarding the pure tones, the Blade Passing Frequency (250Hz) for the power supply block fan is more clearly observed when the block is mounted in the PC casing. When the power supply block is tested outside the casing, the BPF is much less pronounced. Thus, we believe that those high level tones observed when the power supply block is mounted in the PC casing are not exclusively due to an airborne noise propagation path. It is likely that the PC casing is excited by the fan. This shows that the structure-borne noise radiation path can be significant if the PC casing is an efficient radiator in the frequency range of interest.

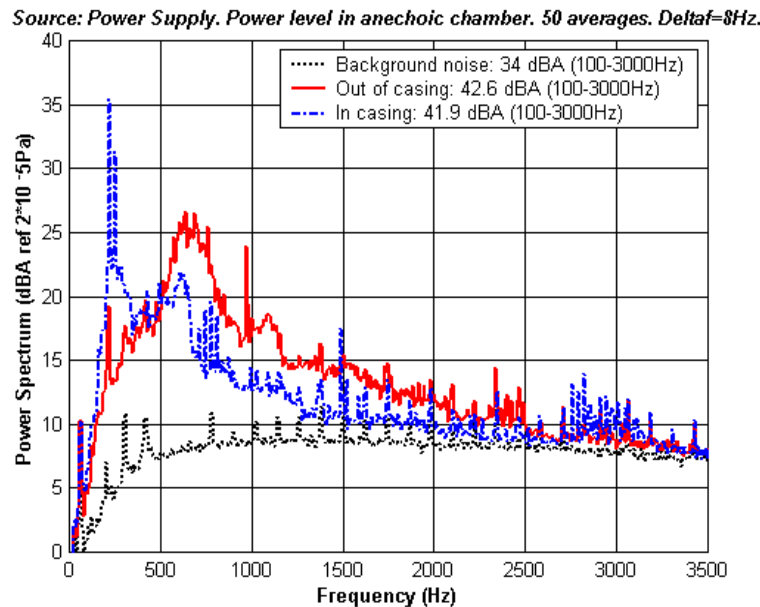


Figure 3.8 Casing effect on power supply block noise (anechoic chamber measurements).

Unfortunately, due to time limitations, we did not further investigate the presence of a structure-borne noise radiation path for the power supply block fan. Structurally decoupling the fan from the power supply block would help to quickly diagnose the importance of the structure-borne path.

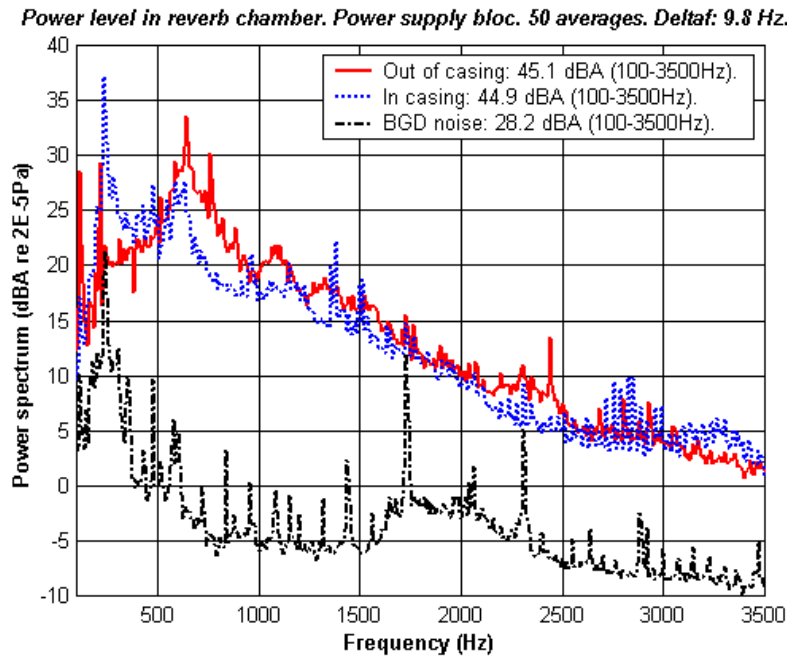


Figure 3.9 Casing effect on power supply block noise (reverb. chamber measurements).

3.2.2.2. Chassis Mounted Processor Fan

a. Location in the PC

In contemporary personal computers, there is usually a fan dedicated to the cooling of the processor fan unit. The fan is often directly mounted on the heat sink connected to the processor. In some instances, it is mounted on the PC chassis, either blowing cool air into the casing (mounted on the front bezel) or blowing hot air out of the casing (mounted on the rear side of the chassis). The latter configuration has been chosen for the Dell computer we are investigating, as can be seen in Figure 3.10.

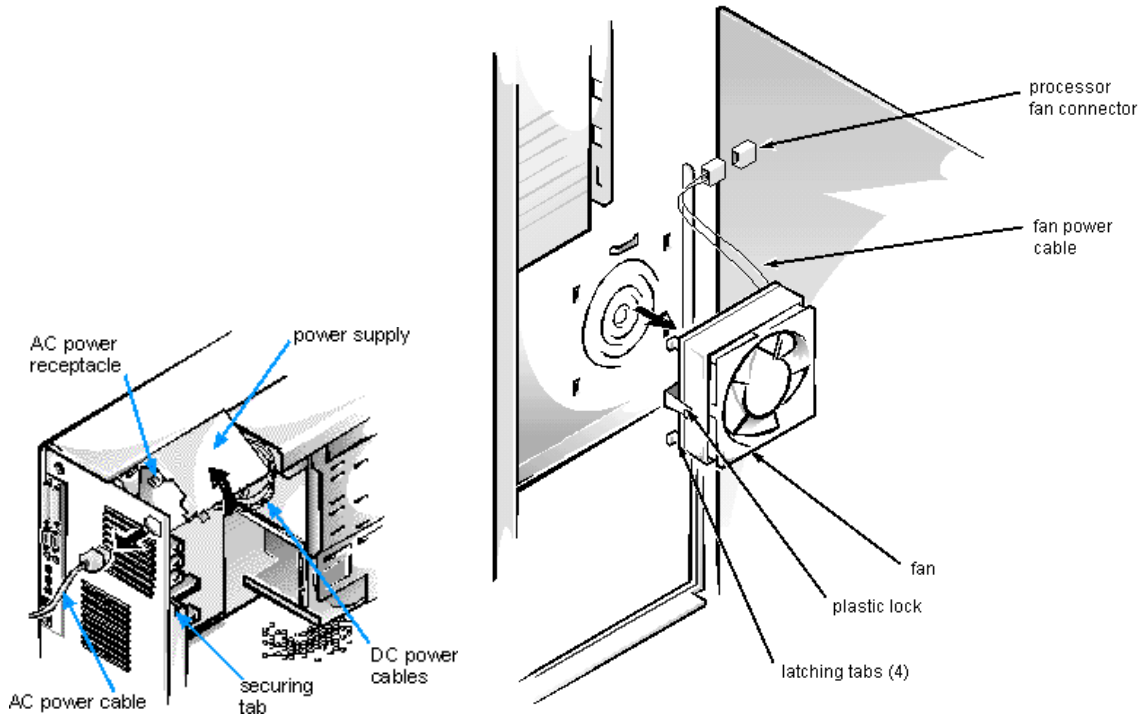


Figure 3.10 Chassis mounted processor fan

A processor fan of 92 mm diameter is located below the power supply block, nearby the processor unit and its dedicated heat sink. The heat generated by the processor is partly expelled out of the casing due to this fan. Similar to the power supply block fan, it is powered by 12V DC voltage and incorporates seven blades. The effective blowing area is approximately 55 cm².

b. Sound Characteristics in PC Casing

The following power level measurements are similar to those performed on the power supply block. The processor fan is powered up by the motherboard, and operates in standard conditions. Nevertheless, both the hard drive and power supply block are powered off to eliminate their noise. An external power supply block, remote located from the measurement room, provides the energy supply to the motherboard by long extension cables. Thus, the processor fan noise is isolated from any other noise PC noise source.

Similarly to the power supply block fan, the sound radiation characteristics of the processor fan vary in time. It will be observed below, that one of the characteristics of the processor fan, when mounted in the PC, is that it radiates multiple tones in the 500-2500Hz frequency range. It was eventually found that the frequency and amplitude of those multiple pure tones would vary as soon as the PC unit was physically manipulated (processor fan unit removed and attached again, movement of the removable covers, etc). Thus, even though the fan unit itself might present fairly stable radiation characteristics in time, its interaction with the PC casing could not be precisely characterized. The second observation concerns the variations among the fan units tested. The sound power level of the processor fan unit, when mounted in the PC casing, is shown in Figure 3.11. The plots correspond to measurements taken at two different dates in a VAL anechoic chamber, with a similar measurement setup (use of ten PCB microphones), but using two different fan units. The blue curve corresponds to the fan originally mounted in the PC donated by Intel. It had been running for hundreds of hours already before being measured. On the other hand, the red curve corresponds to an identical fan model, brand new, that was used for the first time. The brand new fan presents much stronger harmonic content than the used one. The difference is so pronounced that there is a 4dB difference in the overall power level, calculated between 100 and 3000Hz. The differences could be due to variations between the characteristics of the old and new fans. It could also be due to differences in the coupling fan / structure and structure radiation characteristics. The results illustrate the variability likely to be encountered for different PCs of the same model type.

Overall, the processor fan, in its original mounting configuration, mainly radiates pure tones for frequencies between 100 and 2500Hz. Contrary to the power supply block fan, the broadband noise content is low. We believe that, since the processor fan is directly mounted on the casing, the air is flowing with less disturbance than for the power supply. Thus, less air flow induced broadband noise is generated.

Processor Fan in Casing. Power level in anechoic chamber. 80 averages. Deltaf= 9.8Hz.

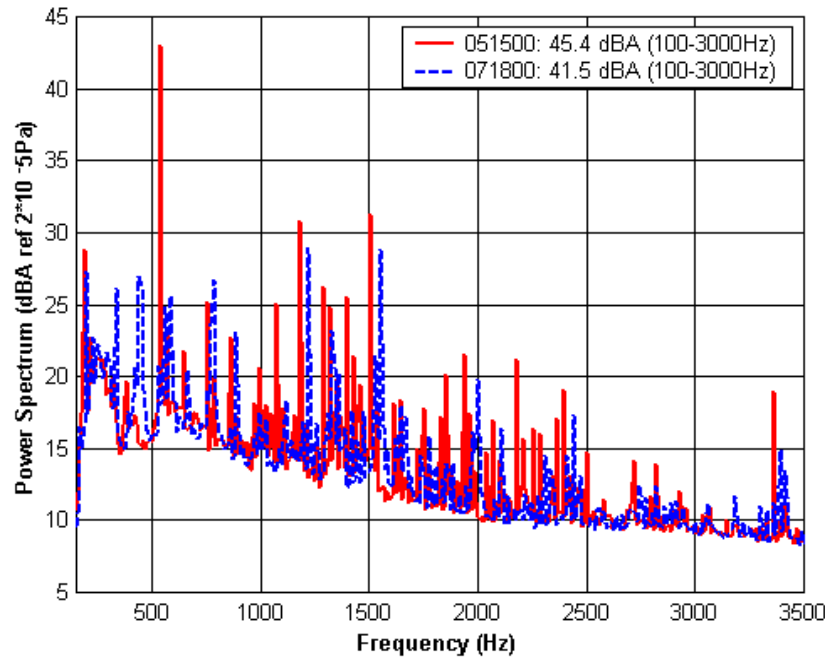


Figure 3.11 Processor fan sound power level. Repeatability issue.

c. PC Casing Effect

The results shown in Figure 3.12 correspond to measurements of the processor fan (in and out of casing (non baffled)) in a VAL reverberant chamber. As said previously, these measurements are not used for absolute sound power determination since no correction factors for the room acoustics were applied.

The results in Figure 3.12 are useful to observe the PC casing effect on the processor fan noise. When the fan is mounted in the casing, the sound power level is higher at low frequencies for both harmonic and broadband noise. This can be due to an increase in airflow induced noise when the fan unit is mounted in the casing. The other explanation is that the fan might present dipole radiation characteristics. Then, the “baffling” effect of the casing might improve the fan acoustic radiation efficiency at low frequencies (relative to the source size), just like for a conventional loudspeaker.

The second observation is the presence of multiple pure tones above 1000Hz when the fan is mounted in the PC casing. This could be partly due to the fact that the power level was measured in the reverberant chamber using several sets of time data (due to the lack of sensors), and that the noise slightly changed characteristics between the acquisitions for the “in casing configuration”. However, the most reasonable explanation is the presence of a structure-borne noise radiation path due to an excitation of the PC casing by the processor fan unit as will be investigated below.

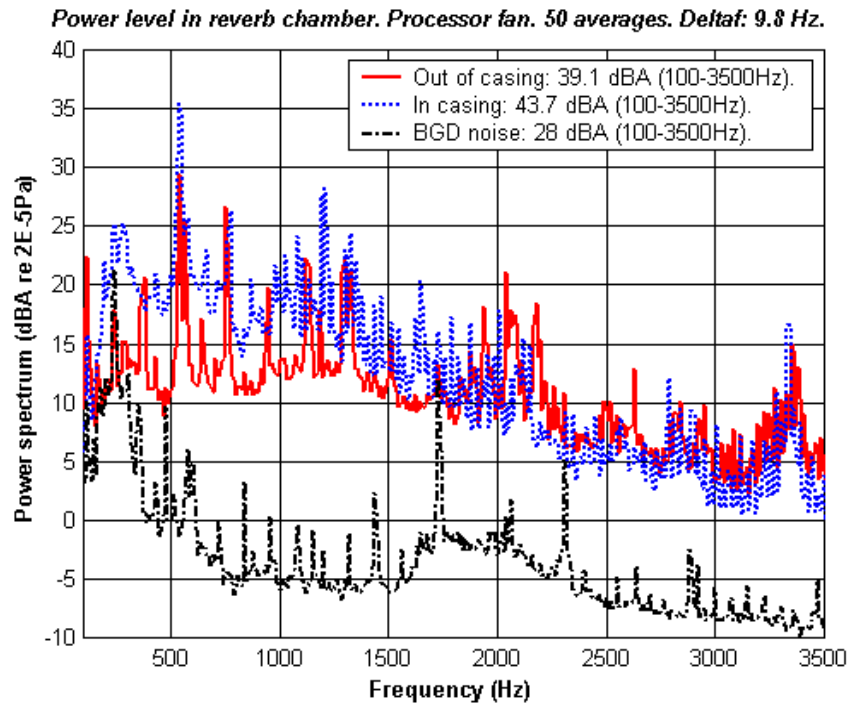


Figure 3.12 Casing effect on the processor fan radiation.

d. Dipole Sound Characteristics

The processor fan is considered an aero-dynamic dipole sound source [25]. In this test, we tried to characterize the noise generated by each of the two monopoles constituting the dipole. Thus, the sound power level of the fan, rigidly fixed in a hole in a baffle, was measured in a half space, as usual with the anechoic chamber method. The measurement setup is illustrated in Figure 3.13.

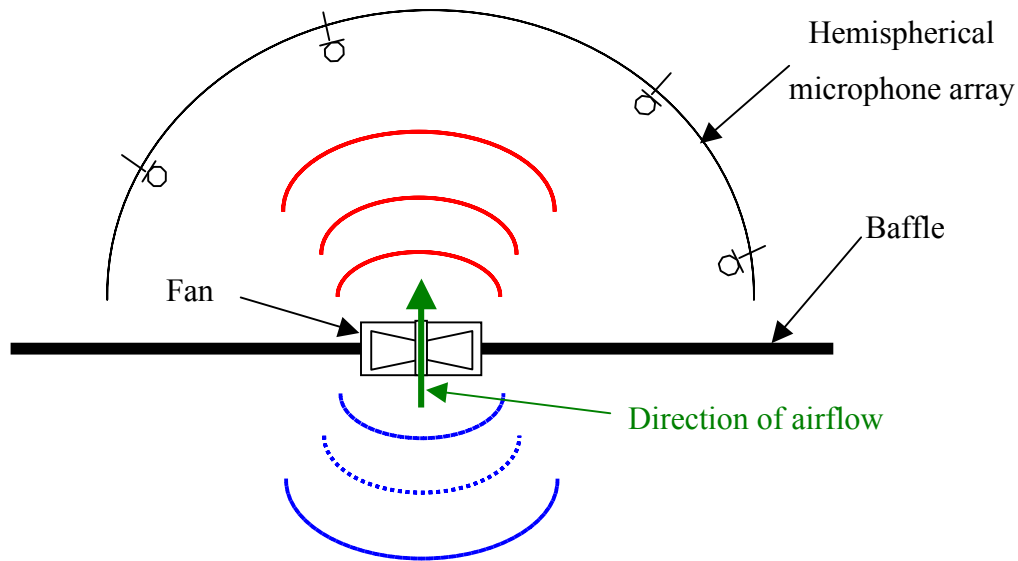


Figure 3.13 Baffled processor fan testing in the anechoic chamber.

Two different results, corresponding to both sides of the processor fan, are obtained as shown in Figure 3.14. The solid red curve corresponds to the power level measured on the half plane where the air is blown into (see Figure 3.13). The dashed blue curve is associated with the half plane where the air is taken from (see Figure 3.13).

In terms of total sound power level, there is just a small variation among the cases (less than 1dB between 100 and 3000Hz). Thus, the source strength is not orientation dependant even though slight differences in the distribution of power are observed in the frequency range of interest. However, these differences are too small to be significant. The fan might have changed radiation characteristics between the two measurements. The baffle might be structurally excited since the fan is rigidly mounted on the baffle. Depending on the mounting, the baffle contribution could be different. Such tests should be performed again with the fan unit structurally decoupled from the baffle.

Source: Baffled Processor Fan. Power level in anechoic chamber. 80 averages.

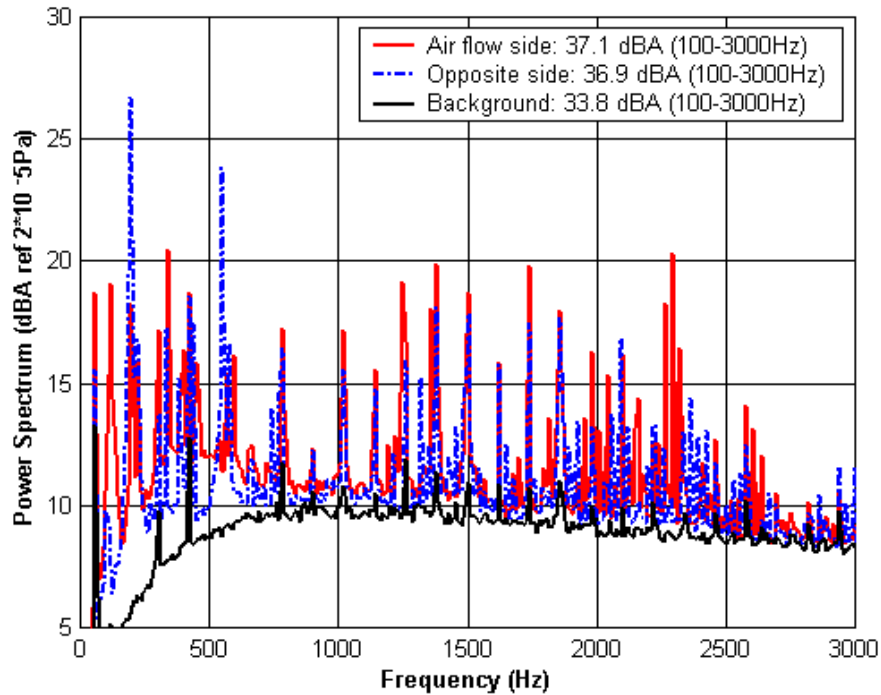


Figure 3.14 Processor fan sound power level. Bipolar characteristics.

e. Protection Screen Effect

As can be seen in Figure 3.15, the chassis mounted processor fan integrates a protection screen preventing a direct access to the fan blades from inside the casing, when this fan unit is attached. The other side of the fan is naturally protected due to its mounting on the rear cover of the chassis.

The same tests as earlier were run, baffling the fan unit in one of the VAL anechoic chambers. Two measurements were performed (fan with / without protection screen), and the results are shown in Figure 3.16. It appears that, when the screen is present, there is an overall increase in the level of the multiple tones radiated, usually around 3dB except for a couple of lower frequency tones (close to the BPF) that are actually reduced in level.

Nevertheless, those conclusions need to be moderated. It was explained earlier that the microphones used for these measurements in anechoic chamber are very sensitive to 60Hz noise. The problem is that, coincidentally, those tones generated by the processor fan are all multiples of 60Hz. The low level fan noise measured is significantly corrupted by measurement noise (see the black curve in Figure 3.16), which includes 60Hz noise and the associated multiple harmonics. Thus, we cannot certify that the conclusions above are reliable. The differences among the different configurations could be partly due to changes in the corrupted content (60Hz noise and harmonics).

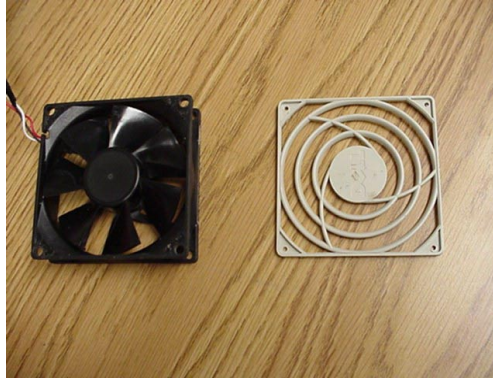


Figure 3.15 Processor fan blades protection.

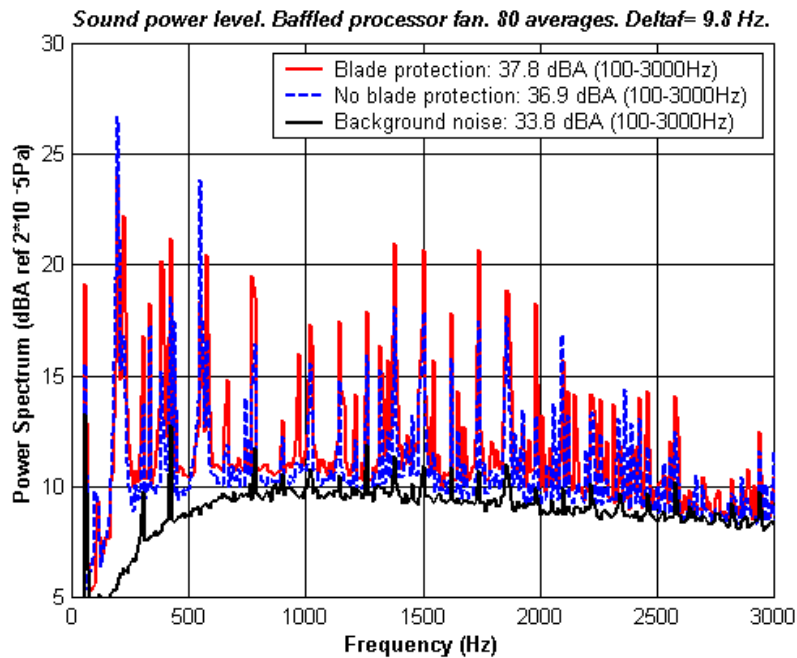


Figure 3.16 Processor fan sound power level. Protection screen effect.

f. Structural Coupling

In this last set of experiments, we want to determine if the noise radiated by the processor fan is mainly airborne, structure-borne, or a mix of both. Since the fan is a vibrating noise source, it might excite the PC casing, which then can radiate acoustic energy. Originally, the processor fan is “snapped” onto the PC chassis with four latching tabs sliding into holes in the metal casing as can be seen in Figure 3.10. In these experiments, the connection attaching the fan to the PC casing was modified. In one case, the objective was to obtain a rigid connection by gluing the fan on the PC casing. In the opposite case, the fan was structurally decoupled from the structure by using soft mounting pads (foam in between the fan and the casing). For each configuration, the sound power level radiated by the system fan/PC casing was measured, and the results are shown in Figure 3.17.

In the top plot of Figure 3.17, the rigid and original mounting configurations are compared. The main tone at 550Hz is reduced by 10dB, which contributes to a large reduction of the overall sound power level (4.5dB). Nevertheless, the spectral distribution of the radiated sound power level remains the same. All the tones, comprised between 700 and 2500Hz, are slightly reduced in amplitude (2 to 5dB) and, more important, shifted in frequency. Looking more carefully at the data, we suspect the processor fan to be spinning at different speeds for the two measurements (changes in power supply characteristics or air flow resistance), which could explain why the tones shift in frequency.

In the bottom plot of Figure 3.17, the soft and original mounting configurations are compared. This time, the major low frequency tone (at 550Hz) is not affected by the new mounting. Thus, the overall power level is not much different from one mounting to the other. Nevertheless, we observe that most of the multiple tones radiated at higher frequencies for the original and rigid mounting cases are now largely attenuated, if not eliminated, by structurally decoupling the processor fan from the PC casing. This important observation suggests that part of the noise radiated by the system processor fan/casing is due to structural excitation of the casing by the processor fan.

The critical frequency of the casing panels is expected to be above 10kHz in the air (thin sheet metal and plastic with not much stiffness [27]). At this frequency, the bending wavespeed in the structure coincides with the sound speed in the propagating medium, which makes the vibrating panel a very efficient sound radiator.

In the present case, the frequency range of structural excitation is well below the estimated critical frequency and should thus not lead to effective acoustic radiation. A possible reason for the significance of the structure borne path could be the presence of many irregularities in the panels as well as stiffeners that increase the panel radiation efficiency. Indeed, below the critical frequency, most of the energy is radiated at the discontinuities of panel (perimeter radiation [27]). Thus, the more discontinuities, the higher the radiation efficiency.

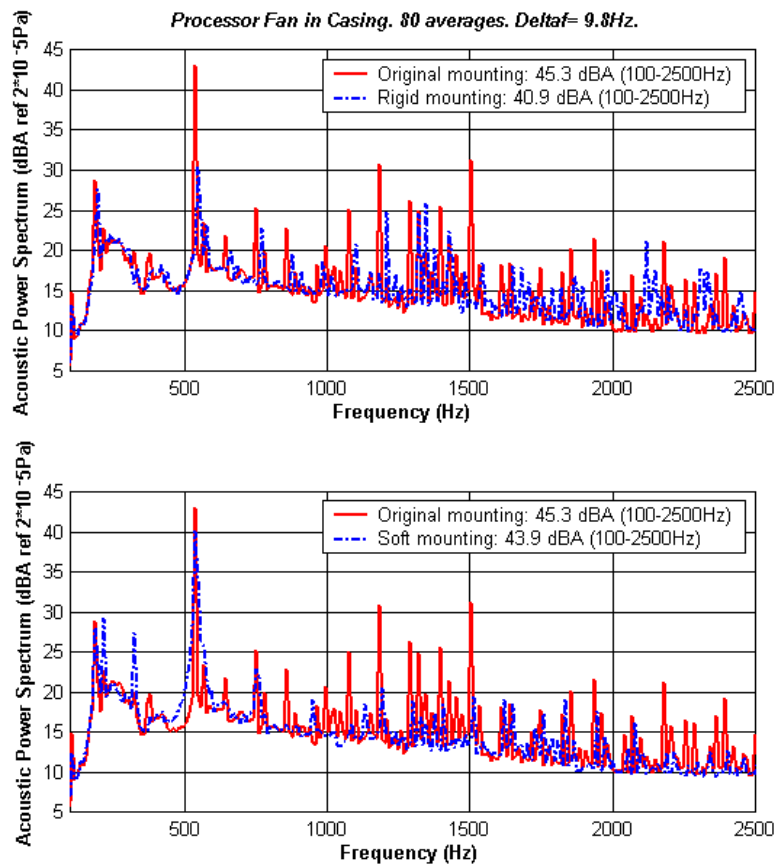


Figure 3.17 Processor fan sound power level. Mounting effect.

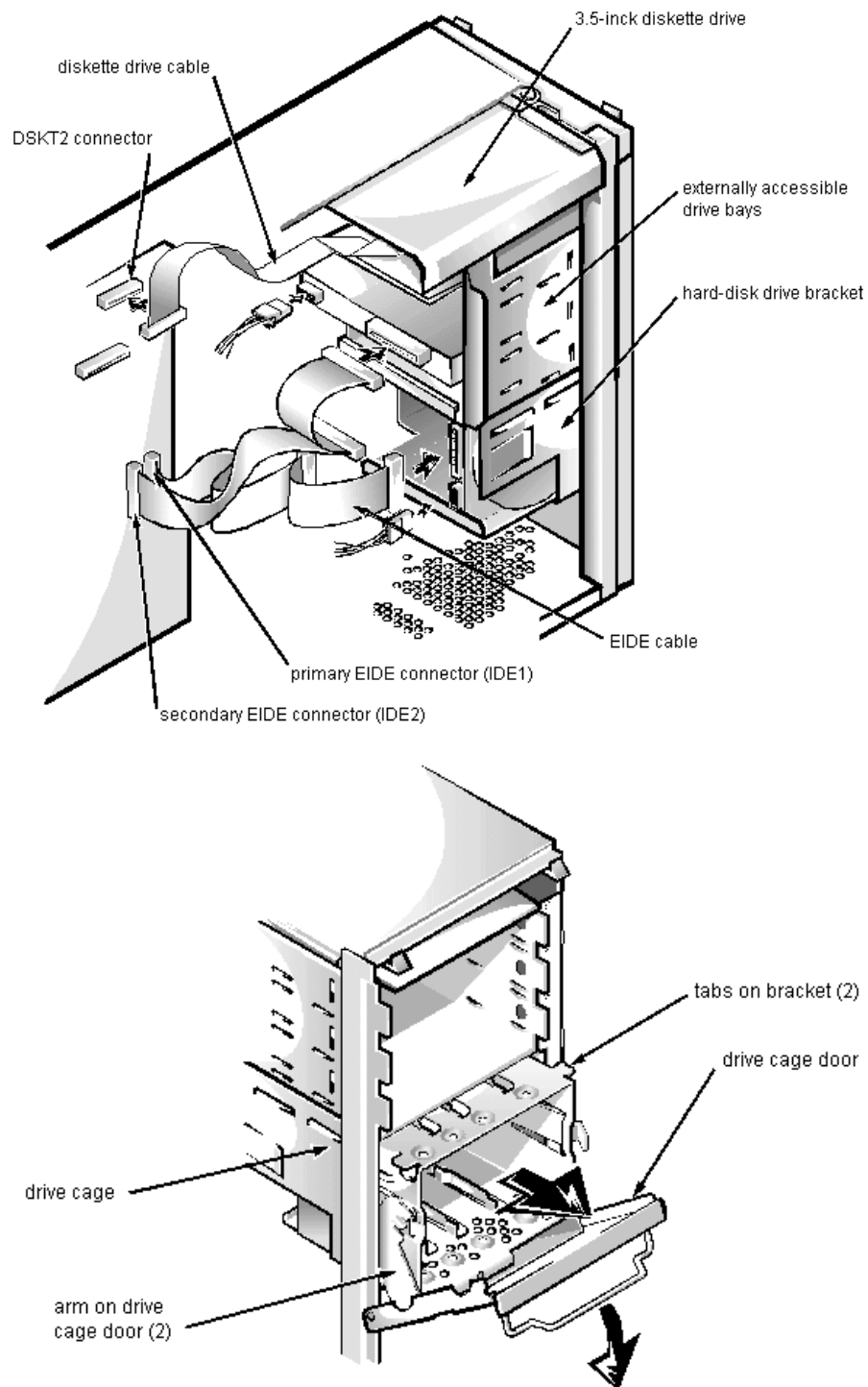
3.2.2.3. Internal Hard Drive

a. Location

This particular Dell computer (Precision Workstation 410) has been provided with a single hard-drive, mounted in the internal bay drive (in contrast with the external bay drive reserved for the CD-Rom, or any device requiring external access). The hard drive studied here is a SCSI model spinning at a nominal speed of 7200RPM.

As can be seen in Figure 3.18, this hard drive is vertically mounted in the internal bay drive. The drive is rigidly attached to the bay with four screws. The bay itself slides in the PC casing, on the front bottom part of the PC casing.

The hard drive is a noise source of interest since it radiates sound continuously while the PC is operating. Two operating modes are distinguished. One is the default mode, when the hard drive is not physically used by the operating system. In such case, the hard drive will be continuously spinning at its nominal speed which is 7200RPM or 120Hz. Since the computer is running under Windows NT environment, there is no power saving mode, thus the hard drive is not powered down at any time when the PC is running. The second mode is called the “seeking mode”. In that case, we hear complex noises with fast transients due to the movement of the heads.

**Figure 3.18 Internal hard drive**

b. Sound Characteristics in PC Casing

Contrary to the other noise sources, the hard drive, in its default operating mode, radiates a fairly stationary noise in the sense that the nominal disc speed is highly stable over time. As described earlier, 60Hz noise can eventually corrupt the measurements. Similar to the processor fan case, this extraneous noise is again difficult to diagnose because the spectral content of the hard drive noise is, coincidentally, also rich in harmonics of 60Hz. The only guarantee that the measurements are “noise-free” is to record the background noise and make sure it is not contaminated by line noise. The second common testing issue is the poor signal to noise ratio as illustrated in Figure 3.19. Except for the tonal content, most of the hard drive noise is dominated by background noise. Finally, Figure 3.19 also illustrates the repeatability issue as the tonal content of the hard drive noise is largely varying in time.

Contrary to the fan noise, the power spectrum of the hard drive, in its original configuration, is dominated by high level pure tones present over a wide range of frequencies: from 500Hz and up to 3500Hz. The three major tones are at 600 (about 5 times the nominal speed of 120Hz), 2100 and 2300Hz. Their amplitude is at least 20dB above the background noise, making them particularly audible.

Active Noise Control does not appear to be the preferred tool to control this particular hard drive since it radiates a significant amount of energy above 1kHz. Active techniques could eventually be investigated for reducing the hard drive noise below 1kHz, and the different radiation paths will be investigated in the next section.

Hard Drive in Casing. Power level in anechoic chamber. 80 averages. Deltaf= 9.8Hz.

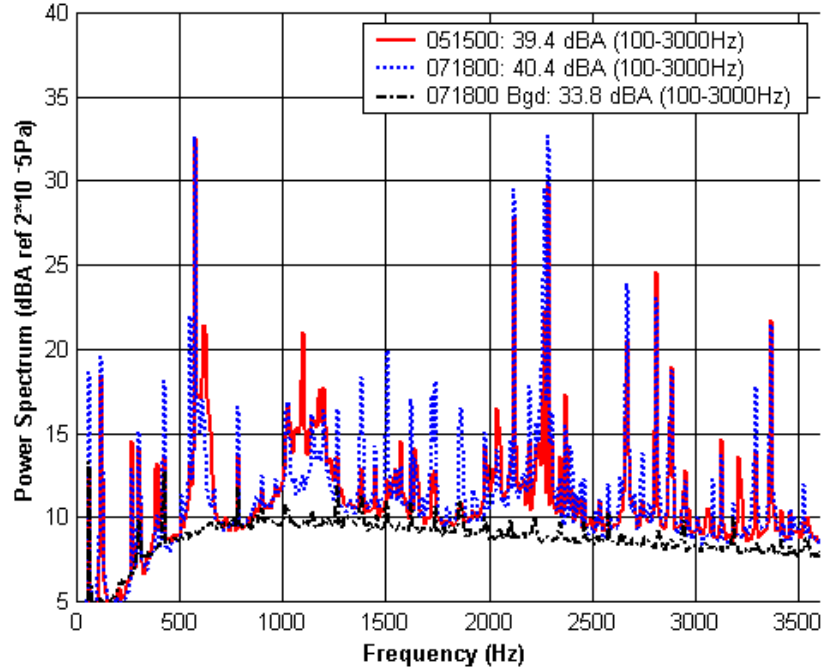


Figure 3.19 Hard drive sound power level. Repeatability issue.

c. PC Casing Effect

The results shown in Figure 3.20 correspond to measurements in a VAL reverberation chamber. The sound power level of the hard drive, while mounted in the PC casing, or in a stand-alone configuration (but still attached to the bracket shown in Figure 3.18) was measured. It can be observed that the low casing transmission loss at high frequencies helps to reduce the hard drive noise (above 2000Hz). On the other hand, below 1500Hz, the casing does not help reduce the hard drive noise, due to either or both:

- A structure-borne radiation path: the PC casing is structurally excited by the hard drive and couples efficiently to the surrounding acoustic medium. This path is investigated in the next section.
- An airborne radiation path: the PC casing may present a high transmission loss at low frequencies, dominated by non-resonant noise transmission (mass law [27]).

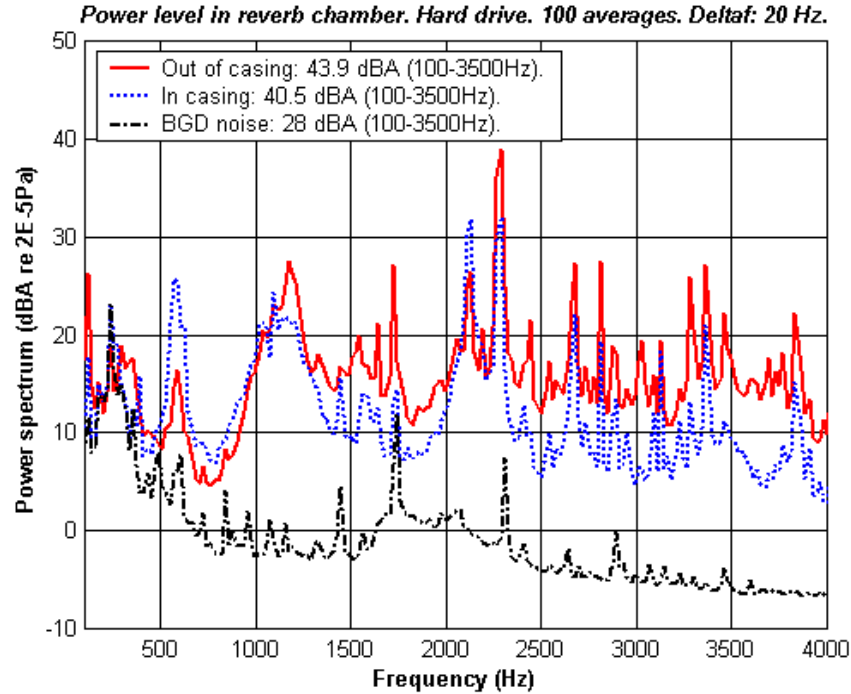


Figure 3.20 Casing effect on the Hard drive radiation.

d. Structural Coupling

For this series of experiments, the effect of the PC casing on the sound power level of the hard drive is studied. The original fixture attaching the hard drive to the PC casing was described as rigid in the introduction. New measurements of the sound power level were performed after structurally decoupling the hard drive from the PC casing. Just like for the processor fan, soft foam pads were used, in this case simply supporting the hard drive at the same location as the original fixture. The results are shown in Figure 3.21.

Results show that the PC casing significantly contributes to the hard drive noise. When the structure-borne radiation path is suppressed, most of the radiated pure tones are attenuated from 6 to 15dB. The overall sound power level integrated in the 100-3500Hz frequency range is reduced by 3.5dB. Thus, there is thus a strong structural coupling between the hard drive and the PC casing. Also, the PC casing appears to be an efficient radiator in the frequency range of interest.

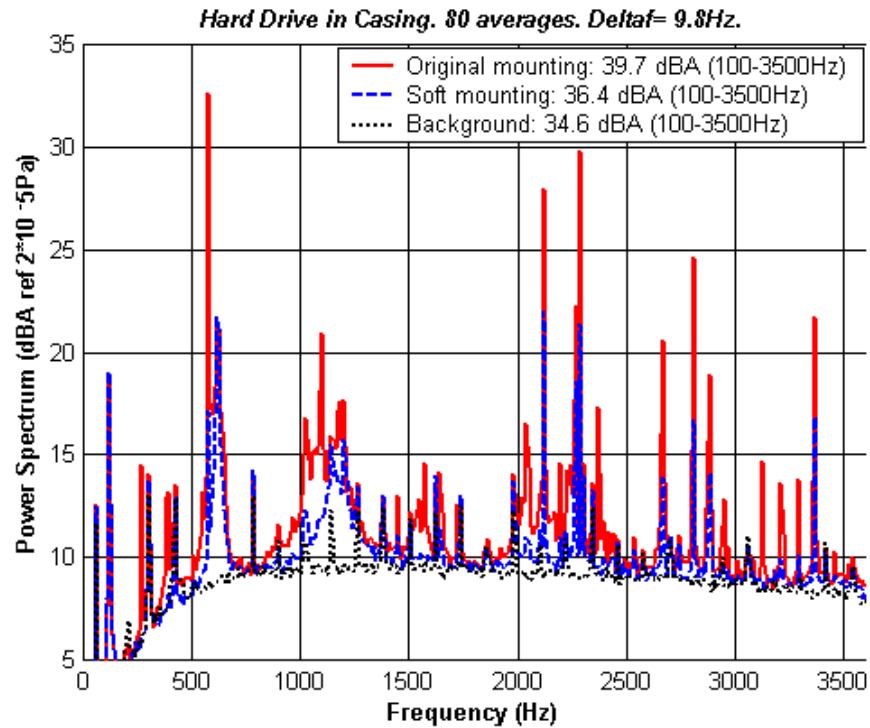


Figure 3.21 Hard drive sound power level. Mounting effect.

3.3. Directivity Survey

Directivity is the characteristic of a sound source to radiate more energy in some directions than others [26]. It is a function of frequency and source complexity. Directivity measurements were performed on the noise sources identified in the PC. After examining the processor fan and power supply block fan results separately, the overall directivity pattern of the PC under study will be studied when all the noise sources are operating.

3.3.1. Experimental Setup and Data Processing

The sound source is placed in a VAL anechoic chamber to simulate free field acoustic conditions. The sound pressure level is measured at different angles from the source at a distance of 1 meter, using ½" B&K microphones (Type 4166) as shown in Figure 3.22. The angular resolution was limited to 10 degrees.

Measurements of the un-baffled processor fan and power supply block (out of the PC casing) were performed in a single plane since those sources present no vertical / horizontal asymmetry. The other measurements were performed on the computer with the three identified noise sources operating. In such case, the directivity was measured in a vertical and horizontal plane for both the front and rear panels.

The signals from the various microphones are acquired in the time domain at a sample rate of 20kHz during a period of 10 seconds using the same acquisition system as for the power level measurements. The acquired time signals are A-weighted using digital filters. Finally, the auto spectra are computed and integrated in one-third octave frequency bands.



Figure 3.22 Directivity measurements in a VAL anechoic chamber.

3.3.2. Results

3.3.2.1. Power Supply Block Fan

The power supply block fan directivity plots for four frequency bands (315, 400, 500, and 1260Hz) are shown in Figure 3.23. A first observation is that the power supply block is not a very directive source since the levels are contained in a 5dB range, from 0 to 180 degrees, at all frequencies. Nevertheless, since the results are integrated over wide frequency bands, the strong directivity patterns of pure tones are smoothed out. The peak levels are usually observed 15 to 30 degrees off the airflow axis. The level is always minimal at 90 degrees except for the 1260Hz third octave band. Overall, the radiation pattern of the power supply fan block seems simple enough to be efficiently actively controlled using simple acoustic sources (like a loudspeaker).

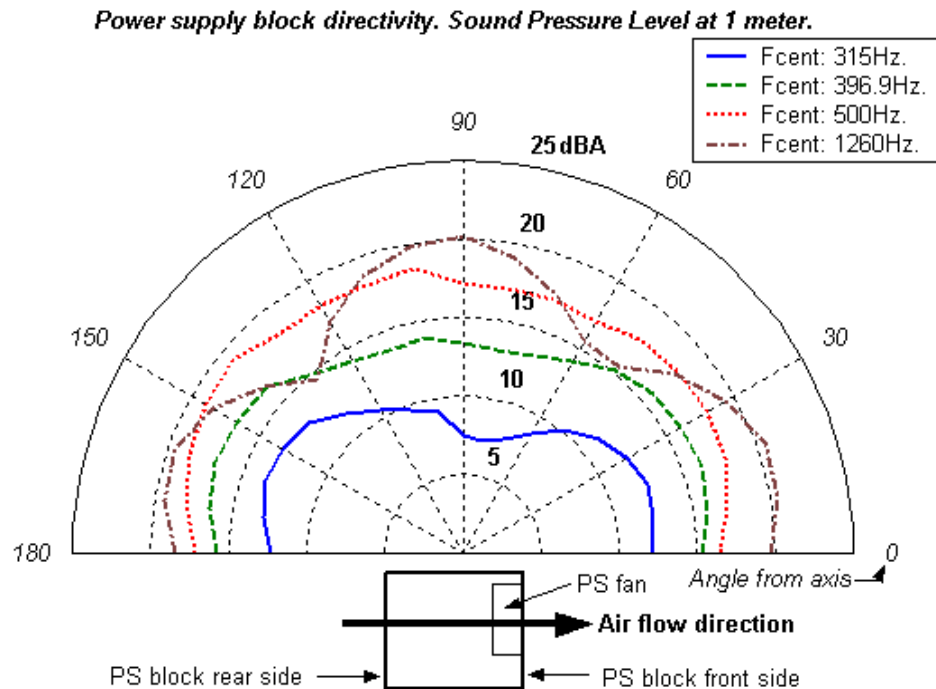


Figure 3.23 Power supply block directivity .

3.3.2.2. Un-baffled Processor Fan

The processor fan directivities for three one-third octave frequency bands (500, 800, and 1260Hz) are shown in Figure 3.24. The un-baffled processor fan appears to be a fairly directive noise source with dipole radiation characteristics. Just like for the power supply block, the peak levels are usually observed 15 to 30 degrees off the airflow axis.

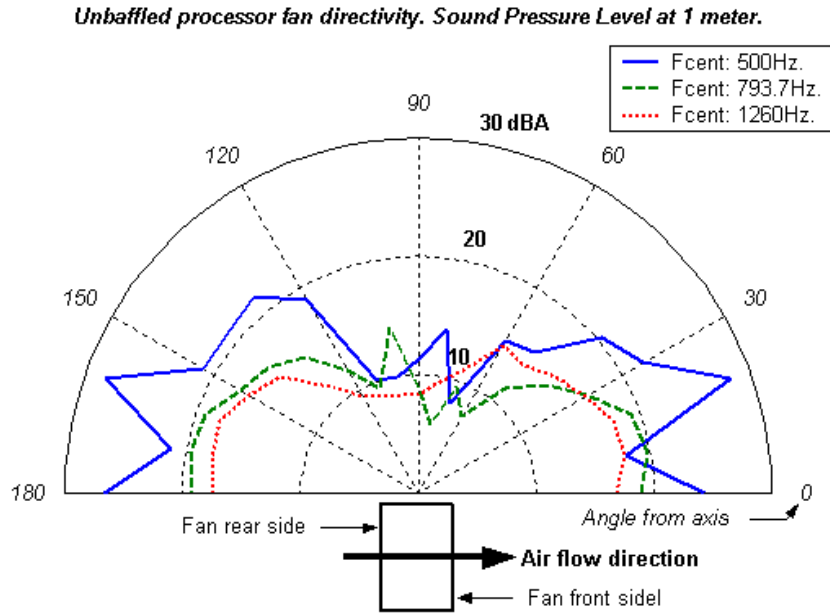


Figure 3.24 Un-baffled processor fan directivity .

3.3.2.3. Overall Results (Sources in Casing)

We now look at the directivity properties of the PC itself when all the noise sources are operating. Figure 3.25 through Figure 3.28 show the directivity in different planes for five third octave frequency bands: 250, 500, 1000, 2000 and 4000Hz.

We observe that the noise radiated by the PC is not very directive up to 2000Hz since the variations in SPL level at 1 meter from the source are usually less than 5dB. The rear panel directivity, in the vertical plane (c.f. Figure 3.25 and Figure 3.26), is similar to what has been observed with the processor fan and power supply block when measured outside the casing: the level is usually maximum 20 to 30 degrees off the flow axis and minimum at 90 degrees (parallel to the fans plane).

If we look at the front panel directivity at low frequencies (250 and 500Hz) in both the horizontal and vertical planes (c.f. Figure 3.25 and Figure 3.27), we observe that the level increases when moving off axis. Since in that frequency range, the rear panel is a major contributor to the overall noise radiated (due the presence of the fans), we believe the above observation illustrates diffraction effects from the PC rear panel. Indeed, on the front panel, the microphones located off-axis might present higher noise levels because they are not only measuring the direct field radiated from the panel but also the “diffracted” sound field from the PC rear panel, where the fans are located.

At high frequencies (2000 and 4000Hz), the directivity plots are not as smooth. It is expected that the sources present strong directivity lobes at high frequencies, but these cannot be observed when the results are integrated in one-third octave frequency bands. Also, the noise is more directive at high frequencies with a level usually maximal in the computer main axis (airflow direction).

To conclude the study of the directivity measurements, the noise radiated by the PC can be considered as not significantly directive. For the PC rear panel, the level is usually higher on the airflow axis probably due to the presence of the fans.

PC front panel vertical directivity. All sources on. Sound Pressure Level at 1 meter.

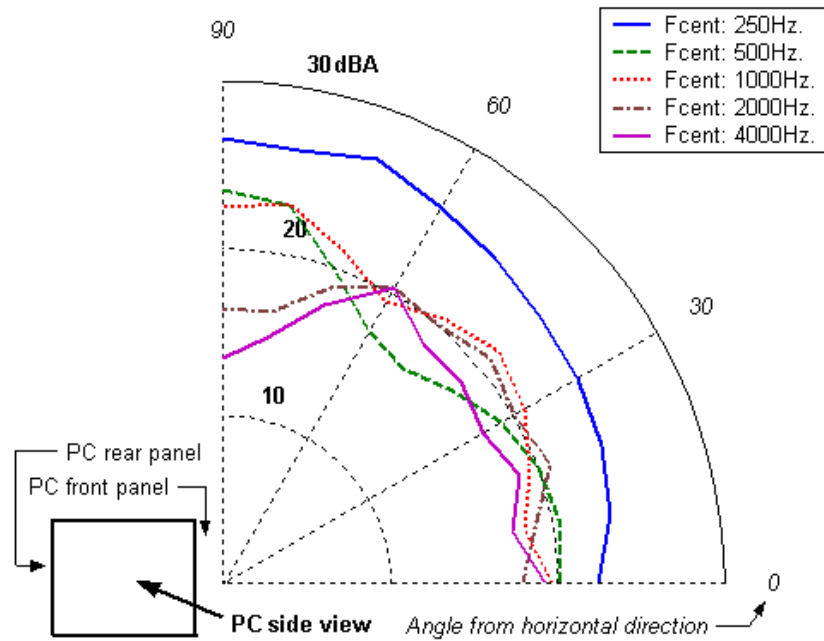


Figure 3.25 PC casing vertical directivity: front panel.

PC rear panel vertical directivity. All sources on. Sound Pressure Level at 1 meter.

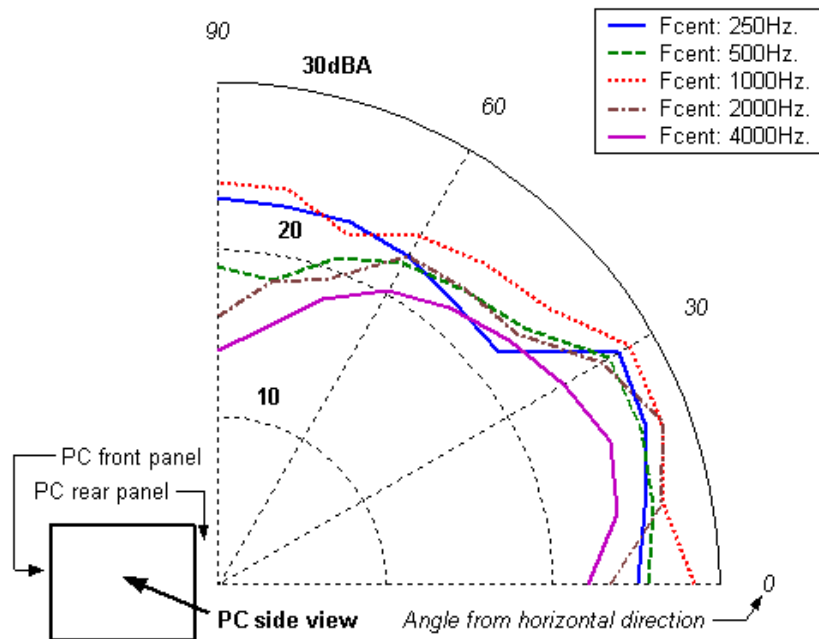


Figure 3.26 PC casing vertical directivity: rear panel.

PC front panel horizontal directivity. All sources on. Sound Pressure Level at 1 meter.

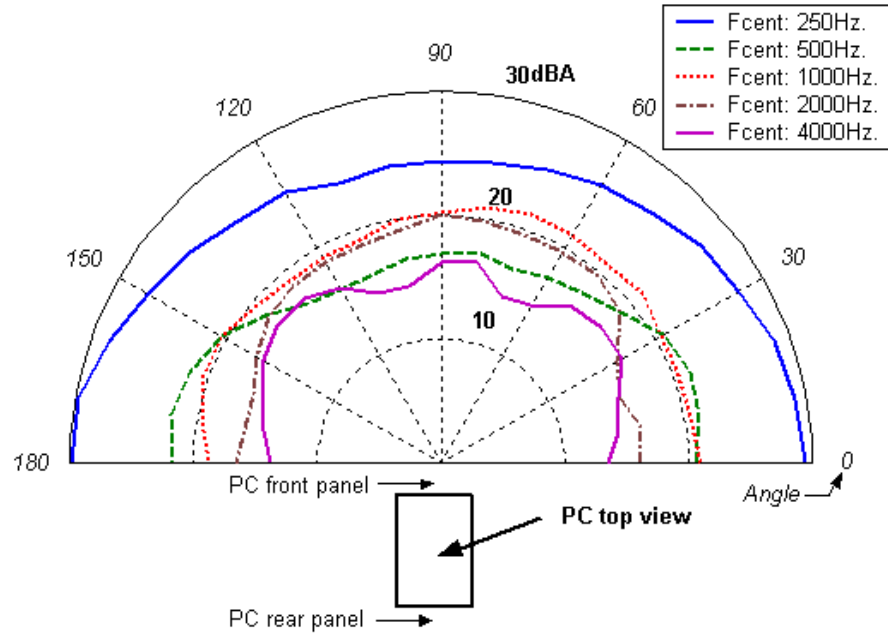


Figure 3.27 PC casing horizontal directivity: front panel.

PC rear panel horizontal directivity. All sources on. Sound Pressure Level at 1 meter.

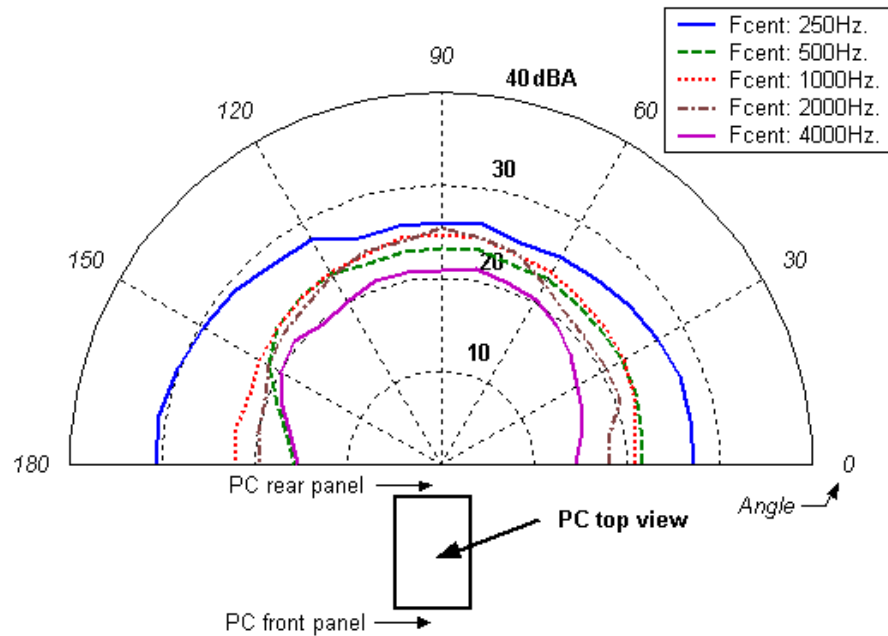


Figure 3.28 PC casing horizontal directivity: rear panel.

3.4. Overall Noise Radiation

3.4.1. Intensity Survey on the Casing

Contrary to sound power level measurements, where the goal was to characterize each noise source, the focus in this section is on sound energy flow radiated out of casing. Sound intensity is a measure of the magnitude and direction of the flow of sound energy. The intensity of a sound field is defined as the averaged sound energy transmitted per unit of time, through a unit area, in a specified direction normal to this surface. The sound power generated by a source must be equal to the normal component of the sound intensity integrated over any surface that completely encloses the source.

In order to measure the intensity, we used a B&K intensity probe using the two-microphones measurement technique [20], which is based on the simultaneous measurement of the sound pressure level by two closely spaced microphones. The intensity is the product of the pressure by the particle velocity (phase related quantities). Thus, by calculating the gradient of pressure between the two microphones, the approximate particle velocity at that location as well as the pressure (mean value of the two microphones pressures) can be deduced. For further details about that measurement technique, one should refer to section 2.2.2.

3.4.1.1. Experimental Setup

For this series of measurements, all the stationary noise sources identified were operating. We performed an intensity scan over some of the PC casing panels including the front and rear panels as well as the side covers. The PC was located in one of the VAL anechoic chambers, on a suspended anechoic floor. The intensity was measured normal to the surfaces at a distance of 10cm using a B&K intensity probe and an automated two-axis traverse as can be seen in Figure 3.29. The surfaces were divided into square areas of 20 by 20mm, leading to 242 measurements on the front and rear side panels and 484 measurements on the side panels. Time histories of the probe microphones were recorded

and latter processed using A-weighting digital filters. The expression used for inferring the intensity from the microphone signals is given below [20]:

$$\hat{I}_r(\omega) = -\frac{1}{2 \cdot \omega \cdot \rho \cdot \Delta r} \cdot \text{Im}\{\hat{G}_{12}(\omega)\}, \quad (3.1)$$

where:

- $\hat{G}_{12}(\omega)$ is the single-sided cross-spectrum function estimate between the two microphone signals.
- ω is the frequency in rad/s
- ρ is density of the medium in which the sound field is propagating (air)
- Δr is the spacing between the microphones in meters (12 mm in our case).

The single-sided cross-spectrum was computed from the time signals based on the expressions developed in section 2.1.1.3. The frequency range of analysis was limited to 100-6500Hz due to inherent limitations in the two-microphones measurement technique. The low frequency limit is due to the small pressure gradient between sensors when the acoustic wavelength is much larger than the probes spacing. The results are then biased when the sensors are not perfectly phased matched. The high-frequency limit is also related to the microphones spacing since the pressure gradient can only be estimated if the acoustic wavelength is at least four times larger than the probes spacing [20]. Since the noise sources present in the PC do not radiate a significant amount of energy above 4000Hz, the latter measurements limitations did not cause problems.

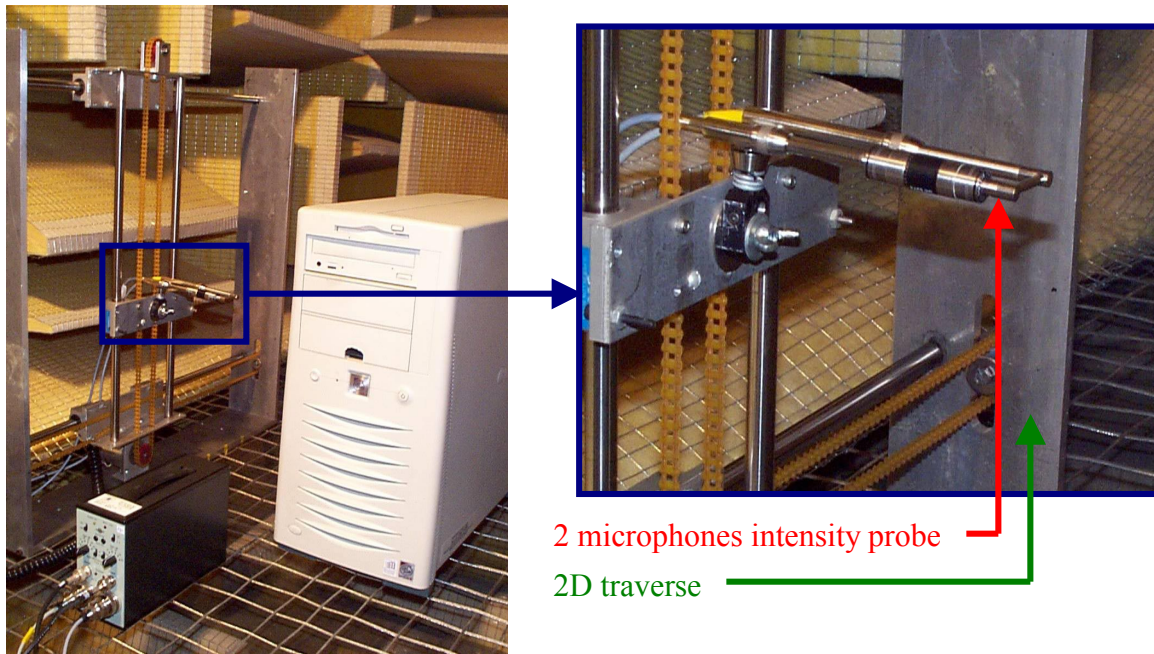


Figure 3.29 Intensity survey on the PC in a VAL anechooic chamber.

3.4.1.2. Results

In Figure 3.30, the power radiated by the four scanned casing panels is shown. This was computed by integrating the intensity results over the measured surfaces. The total power level of 38.6dBA, computed in the frequency range of analysis is about 6dB below the values obtained during specific sound power level measurements (c.f. section 3.2). We believe that this difference is due to the fact that the PC casing was not fully scanned, thus missing the underneath and top panels contribution to the radiation.

The first intensity map shown in Figure 3.31 represents the intensity distribution over the PC casing, in the broad frequency range 100-6500Hz. The fans located on the rear panel constitute the major radiation hot spots, along with the air inlet located on the front panel. We first suspected that the side panels did not contribute much to the sound radiation since no hot spots were apparent. However, when looking at the sound power contribution of each measured panel (c.f. Table 3.1), we observed that these side panels actually radiate more power than the front panel over the frequency range 100-6500Hz.

The intensity is actually not as high on the side panels but it is compensated by the fact that their surface is actually twice as large as the front and rear panel.

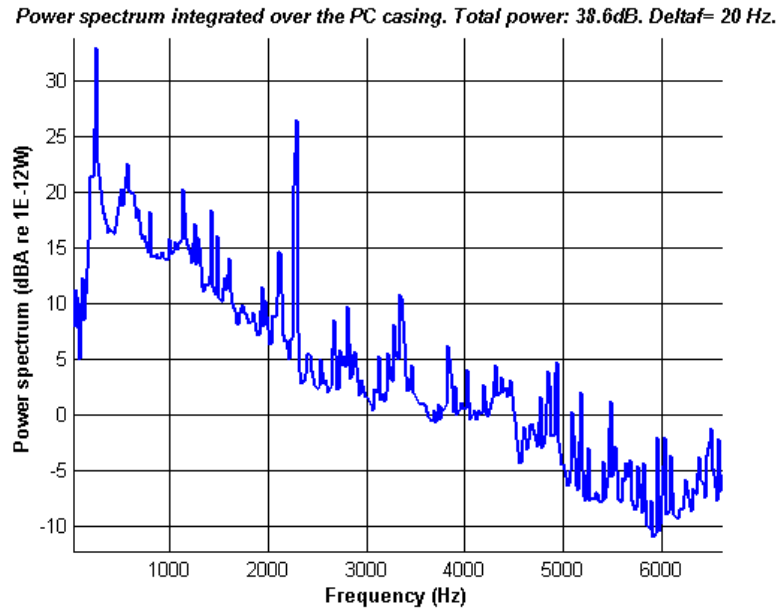


Figure 3.30 Power spectrum integrated over the front, rear and side panels.

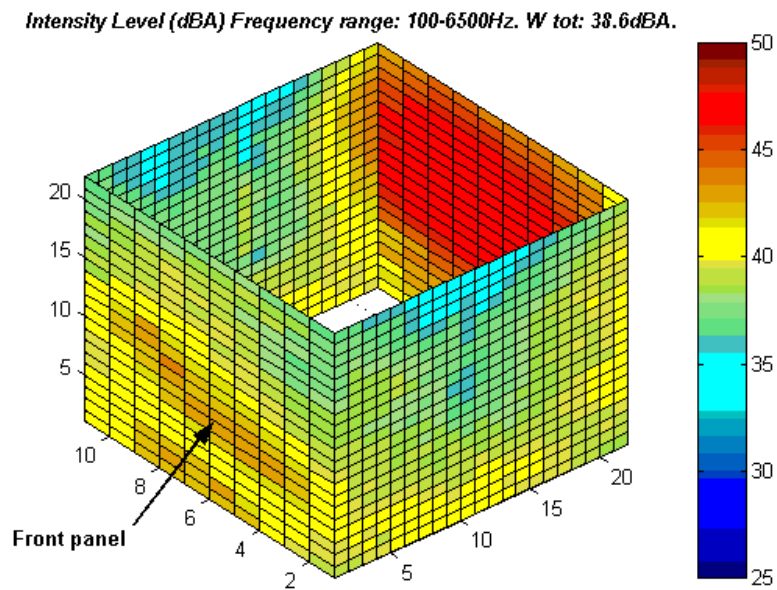


Figure 3.31 Intensity scan over the PC casing. Frequency range: 100-6500Hz.

<i>Central frequency (Hz)</i>	Rear panel Power (dBA)	Front panel power (dBA)	Right panel power (dBA)	Left panel power (dBA)	Total radiated power (dBA)
100-6500	34.8	30.0	31.5	32.4	38.6
250	29.5	23.6	28.3	28.9	34.1
500	23.6	19.2	18.6	20.3	26.9
1260	24.0	17.6	18.8	20.4	27.0
2520	23.2	24.2	18.0	20.0	28.1

Table 3.1 Radiated power distribution over the PC casing measured panels

We can then look at particular one-third octave frequency bands to determine if the previous conclusions do apply at all frequencies. Figure 3.32 through Figure 3.35 represent the intensity distribution in the frequency bands 250, 500, 1260 and 2520Hz. Each of these plots illustrates a characteristic of the sound radiation. The 250Hz band contains mostly the high level pure tone generated by the power supply block fan (Blade Passing Frequency at 245Hz). The power supply block fan outlet remains a radiation hot spot. Nevertheless, as can be seen in Table 3.1, the side panels radiate as much power in that frequency band. We believe that either there is an important noise transmission through the casing at low frequencies (airborne path) or that the power supply block fan is exciting the casing (structure-borne path), which then radiates.

We have to remember that those intensity measurements were taken 10cm away from the PC casing. At such low frequency, we cannot assume to be in the far field of the sound field radiated. Then, these intensity results might only represent near field sound radiation from the PC casing that is evanescent.

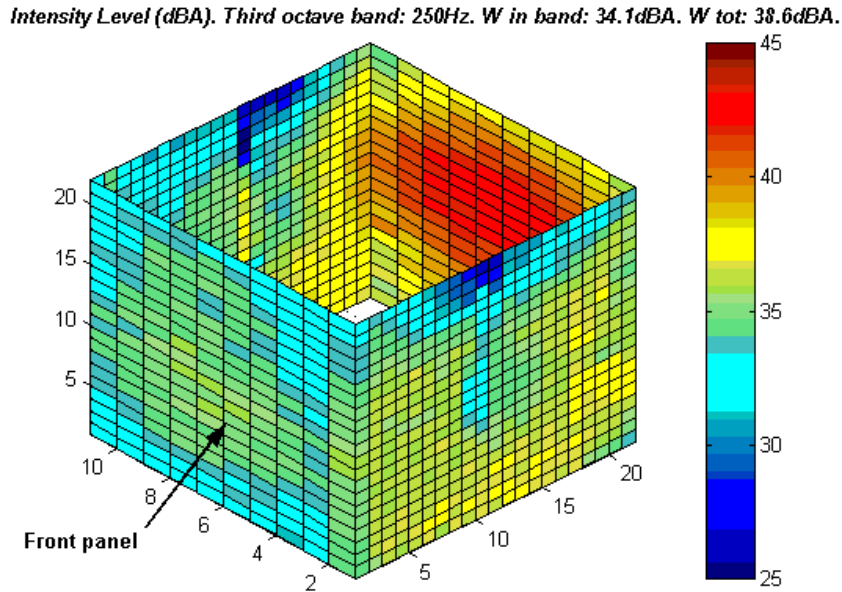


Figure 3.32 Intensity map over the PC casing. Third octave band: 250Hz.

A more typical characteristic of the intensity distribution over the casing when the fan noise is dominating is shown in Figure 3.33 and Figure 3.34. The radiation hot spots are the air inlet and outlet. The intensity level is particularly intense next to the fans. We observe that the higher the frequency, the more focused is the radiation in the air inlet and outlet area.

Finally, the intensity distribution, in a frequency range where the hard drive harmonic tones dominates, is shown in Figure 3.35. The intensity field is complex with strong traces on the front and rear panels not always corresponding to the air inlet or outlet. We will see in Chapter 4 that the high level pure tones generated by the hard drive are mostly due to structural excitation of the casing. This might explain why such a complex intensity distribution is observed on the front and rear panels of the PC, since the metallic casing parts are largely contributing to the overall noise radiation.

Our last comment concerns a limitation of the intensity measurements performed here: the areas above and underneath the PC casing were not scanned. As observed in section 1.1, there are air inlets underneath the PC casing that might represent strong acoustic leaks similar to the air inlet on the front panel.

Intensity Level (dBA). Third octave band: 500Hz. W in band: 26.9dBA. W tot: 38.6dBA.

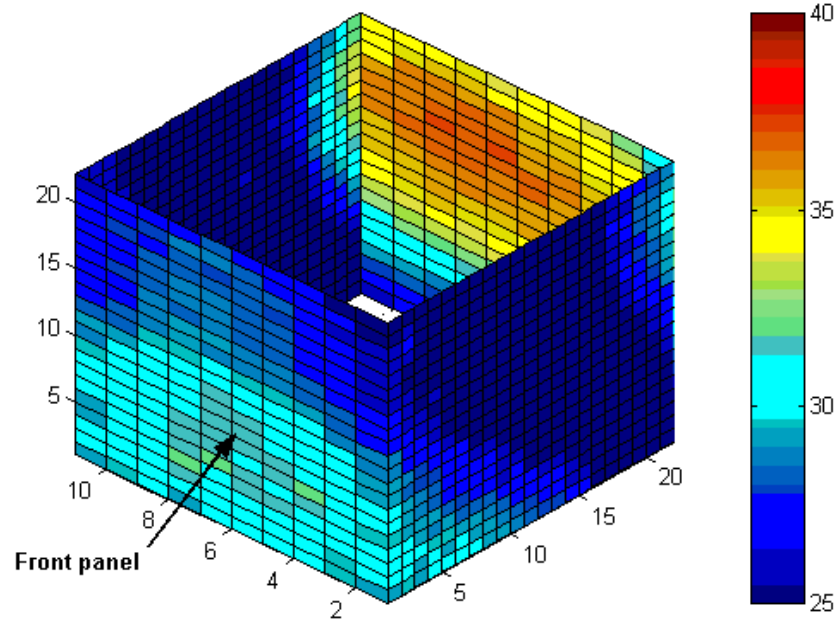


Figure 3.33 Intensity map over the PC casing. Third octave band: 500Hz.

Intensity Level (dBA). Third octave band: 1260Hz. W in band: 27dBA. W tot: 38.6dBA.

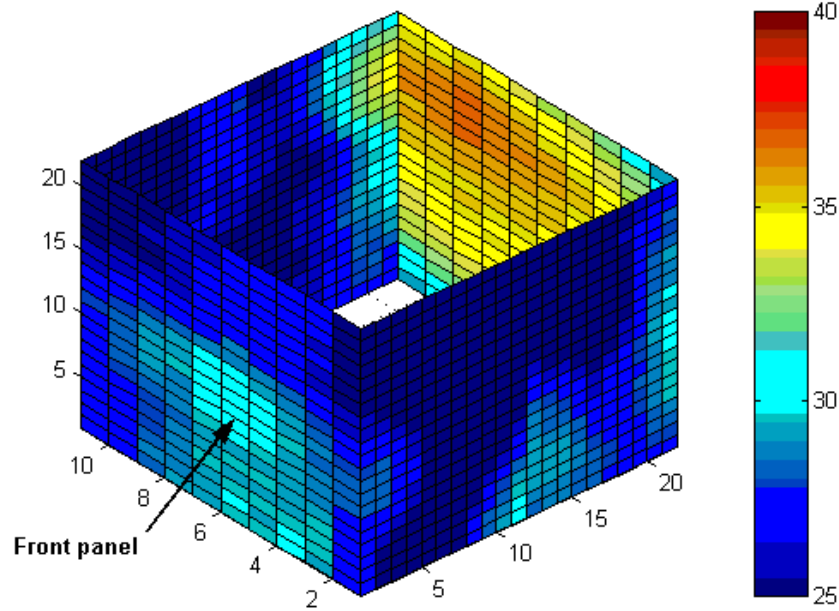


Figure 3.34 Intensity map over the PC casing. Third octave band: 1260Hz.

Intensity Level (dBA). Third octave band: 2520Hz. W in band: 28.1dBA. W tot: 38.6dBA.

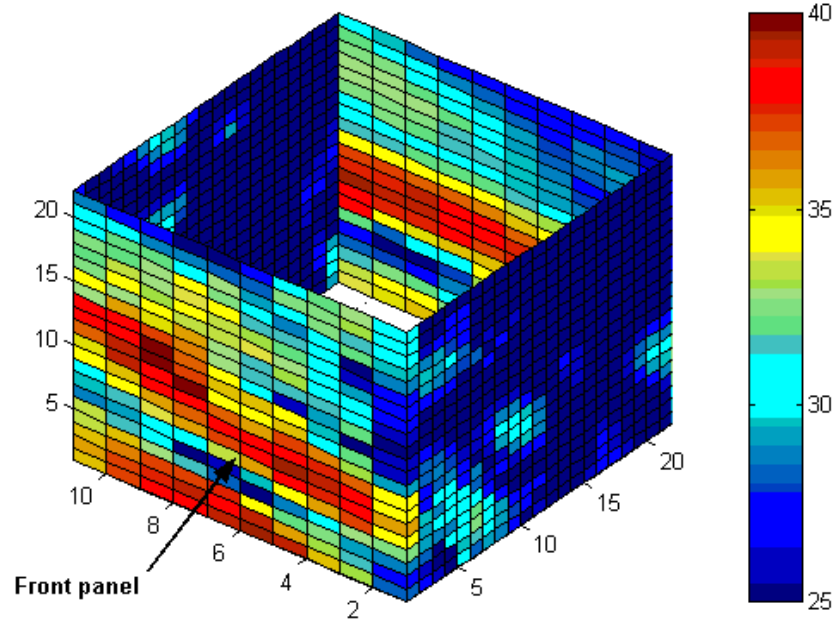


Figure 3.35 Intensity map over the PC casing. Third octave band: 2520Hz.

3.4.2. Ranking of the Noise Sources

Ranking the noise sources in terms of sound power level constitutes a method of comparing the sources regardless of the environment they operate in. The information provided by those measurements helps to determine which source is to be controlled in priority to reduce the overall noise level efficiently. However, in the scope of this project, we also want to control the noise at the head location. In such case, the sources ranking might be different depending upon the operating environment and the directionality of the noise sources. Thus the second objective of the following experiments is to compare the noise sources by measuring the sound pressure level at the head location when the computer is operating in a typical semi-reverberant environment.

3.4.2.1. Sound Power Level

The objective here is to compare the sound power levels of the different noise sources identified in the PC. As observed before, we have to point out the repeatability issue, which makes measurements taken at different times difficult to compare. As illustrated in Figure 3.36, the noise radiated by the PC is not stationary. We believe that some characteristics of the noise radiated are highly dependent on the way some noise sources are mounted in the PC. For example, the processor fan is attached to the PC casing with four latching tabs, as illustrated in Figure 3.10. Such weak connectors may present characteristics varying with time. The fan itself might present significantly different sound characteristics after operating for several hundreds hours for example.

In order to estimate the frequency range of radiation of the PC noise, it was placed in a VAL anechoic chamber and the sound pressure level was measured at one meter on the front axis using a high precision B&K microphone. As can be observed in Figure 3.37, the noise sources identified do not radiate above 6000Hz. As we have seen before, most of the radiated noise is below 3000Hz. The high level tones observed here between 4000 and 6000Hz are due to the hard drive and present a strong directivity pattern since the overall sound power radiated in this frequency range is very small.

In terms of source contribution, Figure 3.38 shows that the power supply block fan is a major noise source below 500Hz for both broadband and pure tone radiation (top plot). The processor fan is mainly contributing to the noise radiated between 500 and 1500Hz (middle plot). Some strong tones between 1100 and 1300Hz are most probably due to the processor fan, but this cannot be easily observed due to the repeatability issue. The hard drive is only responsible for the pure tones radiated between 1500 and 3000Hz (lower plot). The pure tone at 600Hz does only slightly contribute to the overall sound power level.

All sources in Casing. Power level in anechoic chamber. 100 averages. Deltaf: 20 Hz.

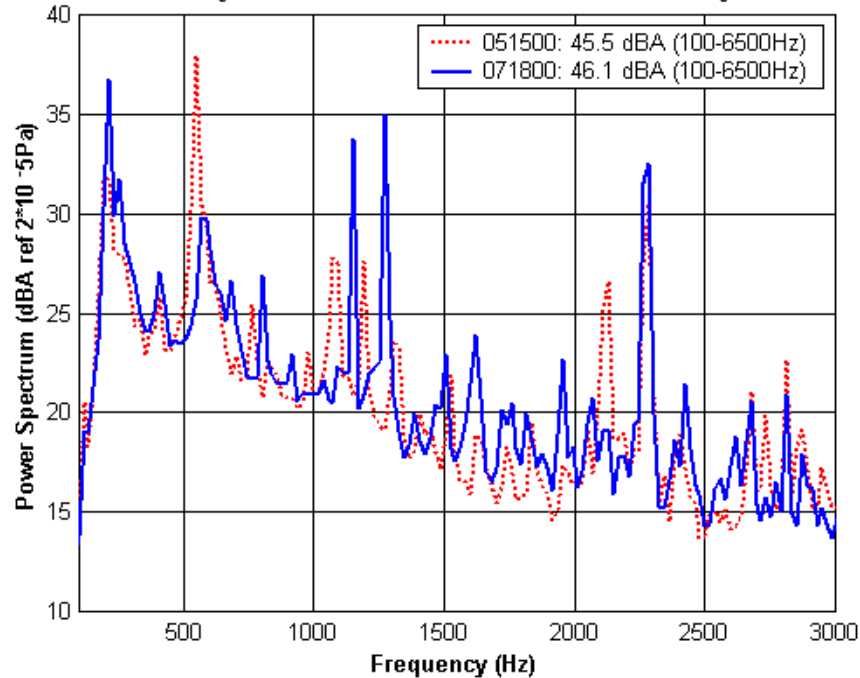


Figure 3.36 PC overall sound power level. Repeatability issue.

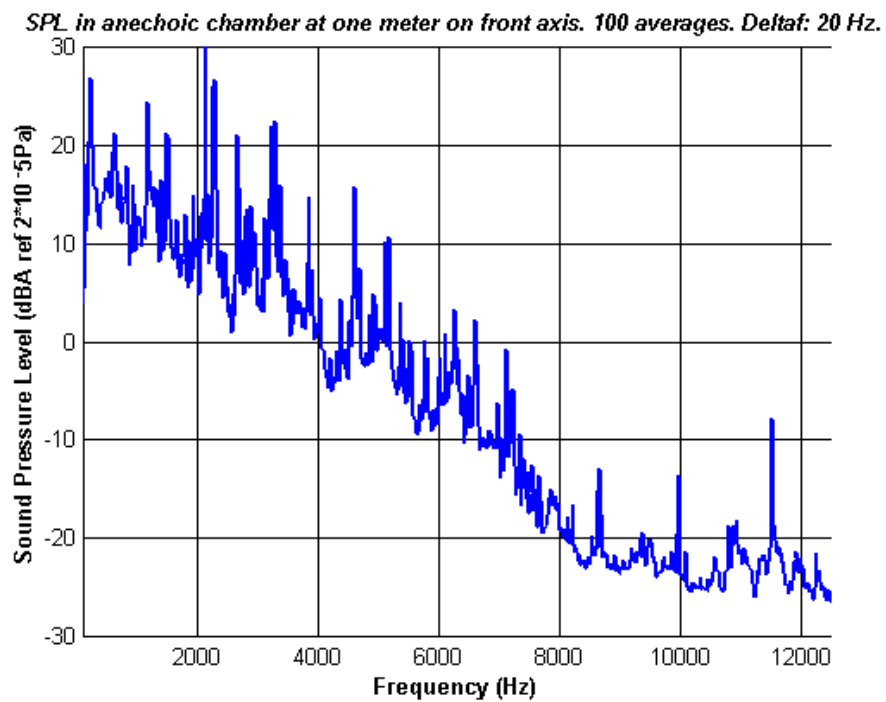


Figure 3.37 PC overall sound power level. Broad frequency range.

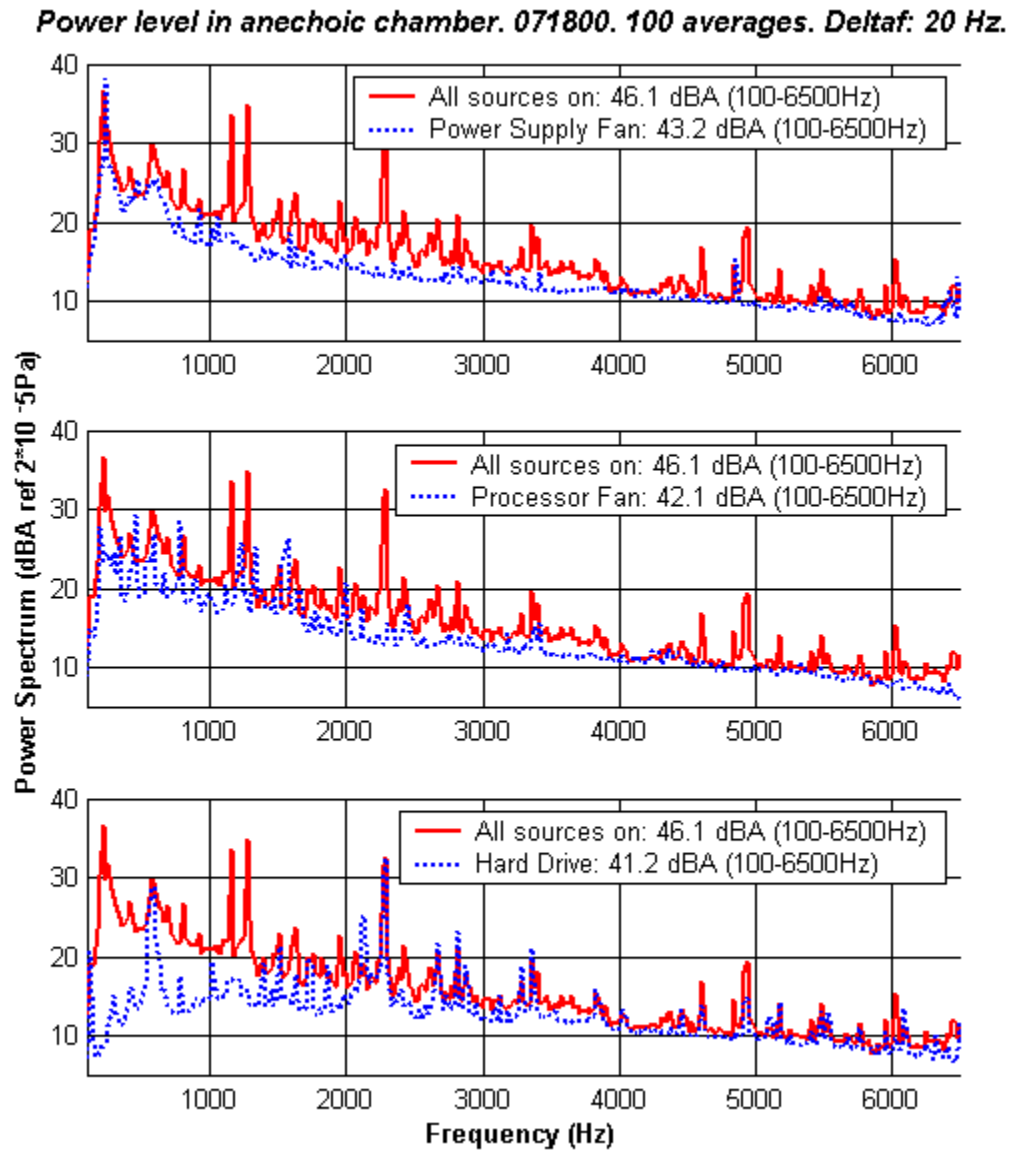


Figure 3.38 PC overall sound power level. Broad frequency range.

The frequency range of dominant sound radiation can be clearly distinguished for the different noise sources. The objective is not to quantify the difference in sound power level radiated. Thus, the sound power level of each noise source is computed in the three frequency bands illustrated earlier as can be seen in Table 3.2. Due to repeatability issues, the results below could vary from +/- 0.5dB depending upon the measurement conditions.

Frequency range:	BDG (dBA)	PS (dBA)	PF (dBA)	HD (dBA)	All (dBA)
100-500Hz	24.0	40.5	36.5	26.5	41.0
500-1500Hz	30.0	37.0	38.0	36.0	42.5
1500-3000Hz	31.5	33.0	35.5	38.0	39.0
100-3000Hz	34.0	42.5	41.5	40.5	45.5

Table 3.2 Sound power level of the PC sources.

The power supply block fan is the noisiest sound source in the PC with an overall sound power level of 42.5dBA in the 100-3000Hz frequency range. This source is also dominating the radiation below 500Hz, which makes it an ideal candidate for Active Noise Control.

The processor fan comes second with an overall sound power level of 41.5dBA. It is generating noise over a wide frequency range with a slight dominance between 500 and 1500Hz. This sound source should be controlled by combining passive / active techniques.

Finally, the hard drive is the quietest source with an overall level of 40.5dBA. It does mainly radiate sound between 1500 and 3000Hz, and also less significantly below 1500Hz. Such a noise source should be controlled using passive techniques exclusively.

As a general observation, we will note that the sum of the sound power level radiated by each source separately, is higher than the actual sound power level of the three noise sources operating simultaneously (more than one dB difference). This may be due to some interaction among the three sound fields at frequencies where they all radiate harmonic sounds.

3.4.2.2. Sound Pressure Level at the Head Location

We have seen in the previous section that the computer used for this project is not significantly noisy with an overall sound power level around 45dBA. Consequently, any acoustical measurement has to be performed in an anechoic chamber so that the acoustic background noise is not an issue. For this series of measurements, the computer was located in one of the VAL anechoic chambers. In order to simulate a simplified user environment, the PC was placed on a reflective floor (wooden plate with painting) as can be seen in Figure 3.39. One other reflective wood panel was placed behind the computer, acting as a rear wall. Finally, a desk was located next to the PC to better simulate a typical user environment. The microphone was at the approximate height of 1.5 meter corresponding to the ears location for a typical user sitting in a chair. Such a setup was also used for the local active control of noise that will be discussed in Chapter 6.

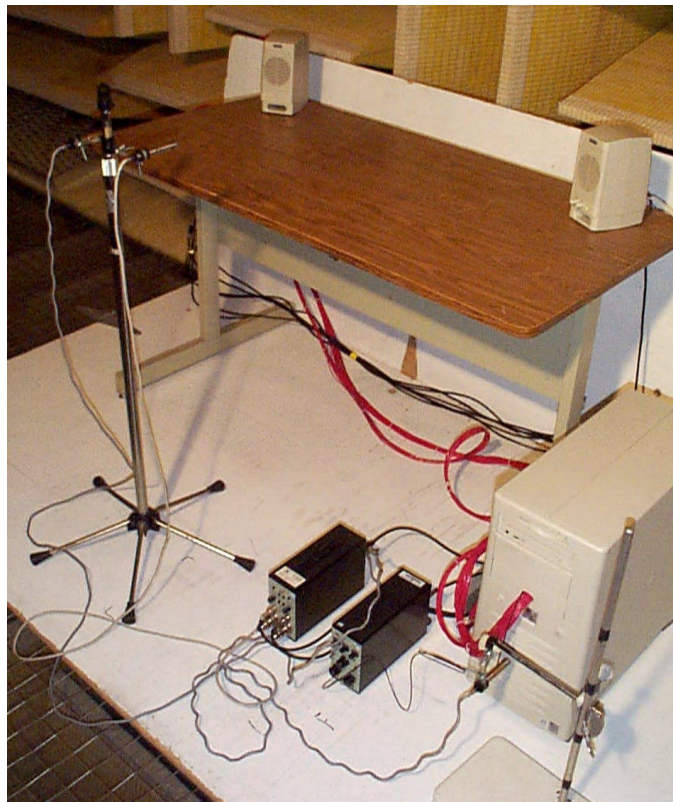


Figure 3.39 User environment simulation in a VAL anechoic chamber.

Similarly to the analysis performed with the sound power level measurements, we tried to evaluate the contribution of each noise source to the sound pressure level at the head location. Typical results for one specific microphone location are shown in Figure 3.40 and Table 3.3 where the overall SPL levels in specific frequency ranges are compared.

Contrary to before, the major contributor over the 100-3500Hz frequency range is not the power supply block fan but the processor fan, as shown in Figure 3.40. The different noise sources are still operating in different frequency ranges. The power supply block fan remains a favorite candidate for Active Noise Control (top plot). On the other hand, both the processor fan and hard drive will require passive redesign in order to improve sound performance at higher frequencies since the Sound Pressure Level is quite high above 1000Hz and up to 3500Hz (middle and lower plots).

Frequency range:	PS (dBA)	PF (dBA)	HD (dBA)	All (dBA)
100-600Hz	25.6	21.0	16.8	28.3
600-2200Hz	25.5	28.7	21.2	31.1
2200-3500Hz	13.0	16.2	21.1	22.0
100-3500Hz	28.4	29.5	24.8	33.2

Table 3.3 Total sound pressure level at the head location (user environment).

SPL at head location. Simulated user environment. 100 averages. Deltaf: 20 Hz.

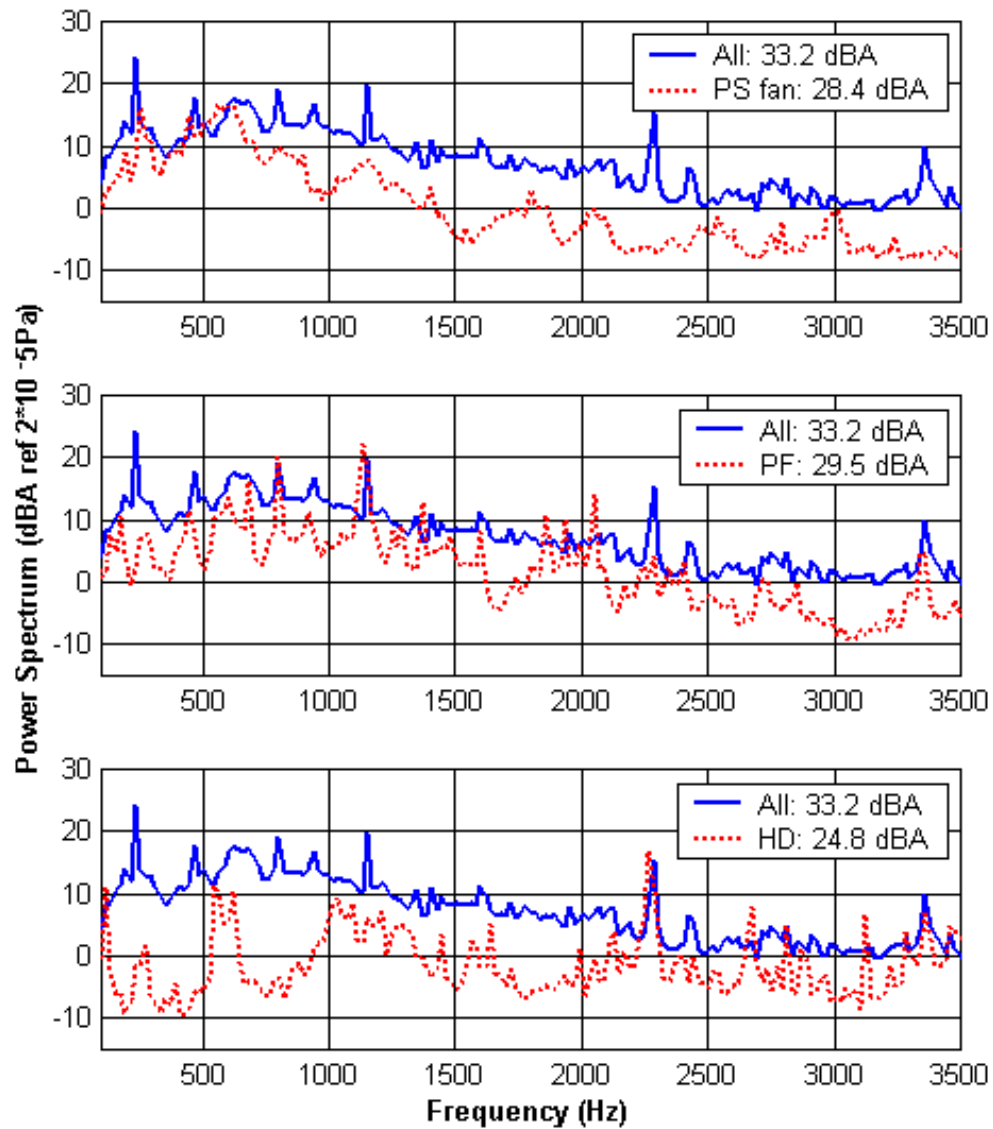


Figure 3.40 SPL at head location. Simulated user environment.

Chapter 4. Passive Control of PC Noise

In Chapter 3, the different stationary noise sources present in the PC were identified and characterized. It was observed that the PC radiates noise over a wide frequency range, from 200Hz up to 6000Hz. One important conclusion was that the noisiest source, which is the power supply block with its integrated fan, radiates most of its energy below 1000Hz. It is thus an ideal candidate for Active Noise Control. Nevertheless, we have also observed that the two other noise sources generate significant noise levels above 1500Hz. In such case, a passive treatment needs to be developed, which is the subject of this chapter.

The structure-borne noise radiation path in the PC constitutes one problem we tried to solve using passive techniques. The idea was to structurally decouple the vibrating primary noise sources from the casing unit. Part of the airborne noise radiation path was to be controlled passively using lined ducts. Those ducts were also designed to permit the integration of Active Noise Control technology so that a global control of PC noise could be achieved over a wide frequency range (below and above 1000Hz). This chapter is limited to the study of the passive noise control treatments we implemented in the PC.

4.1. Structural Decoupling of Vibrating Noise Sources

4.1.1. Processor Fan Noise

In section 3.2.2.2, we have observed that the noise radiated by the processor fan mounted in the PC casing was a combination of airborne and structure-borne radiation paths. In order to limit the efficiency of the structure-borne radiation path, the mounting of the processor fan on the casing was modified. The original connection system is shown in Figure 4.1.

We believed at first that such a weak connector was creating some rattle effect that appeared in the form of multiple tones in the power spectrum when the fan was mounted in the casing. Nevertheless, rigidifying the connection did not help to attenuate the high frequencies tones. This previous hypothesis was then excluded.

The second explanation on the origins of those multiple tones was that the processor fan was exciting the PC casing, which then was radiating noise. By mounting the fan on the PC casing using glue and foam, we could structurally decouple the vibrating source from the PC casing. The results obtained showed a significant reduction in the amplitude of the tones radiated between 1000 and 2500Hz as can be seen in Figure 4.2. Even though the fan BPF (around 250Hz) has a higher level when the fan is soft mounted on the casing, an attenuation of almost 2dBA is still obtained over the 100-2500Hz frequency range. The power spectra shown here are slightly different from those in Chapter 3. This is because the measurements were taken at two different dates and the noise characteristics of the noise sources are not stationary as explained earlier. Nevertheless, the structural decoupling of the fan revealed effective in both cases with an attenuation of the pure tones of up to 15dB.

After modifications, the noise radiated is still rich in pure tones as can be seen in Figure 4.2. However, many of the tones might be due to the presence of measurement noise (60Hz and its harmonics). In particular, the background noise shows that the signals are corrupted.

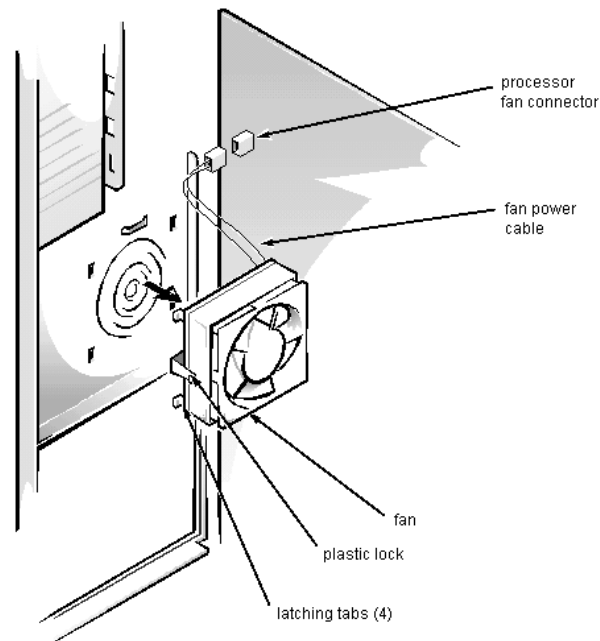


Figure 4.1 Chassis mounted processor fan

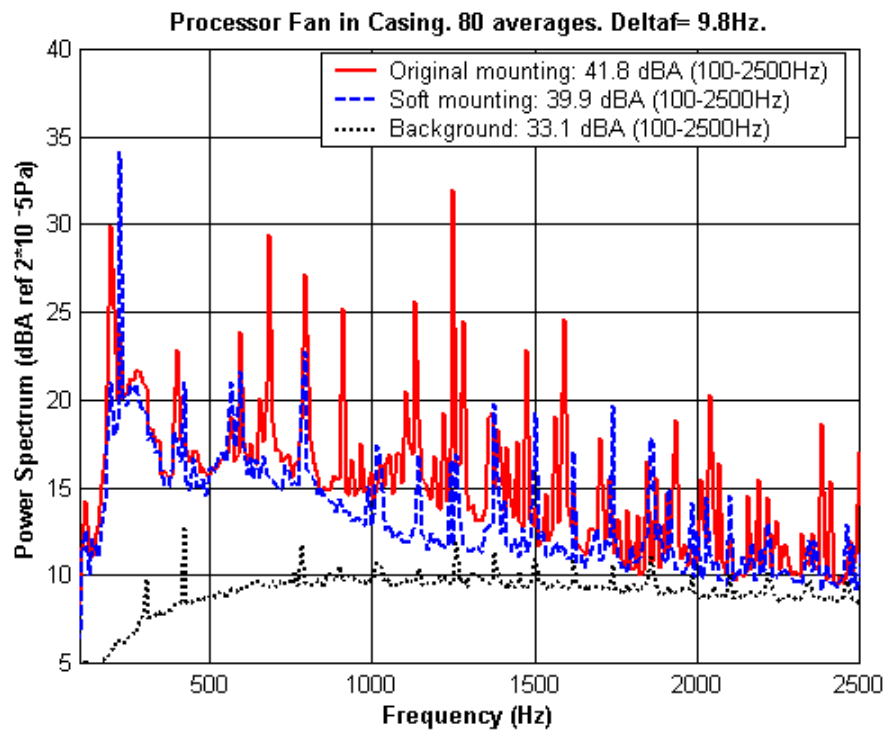


Figure 4.2 Processor fan structural decoupling effect on PC sound power level.

4.1.2. Hard Drive Noise

We also identified a structure-borne component to the hard drive noise by altering the connection between the unit and the PC casing. Instead of rigidly mounting the hard drive vertically in the dedicated bracket as can be seen in Figure 4.4, it was mounted horizontally. The hard drive was then simply supported by the bracket at three locations by soft “suspensions” made of foam, thus vibration isolated from the PC casing.

The comparisons between original and soft mounting are shown in Figure 4.3. All the pure tones radiated by the hard drive are attenuated by up to 15dB when structurally decoupled from the PC. The overall reduction in the sound power level of the PC, with the hard drive as unique operating source, is around 3.5dBA between 100Hz and 3500Hz. Similarly to the processor fan, the airborne noise radiation path is not affected by this control treatment. This is the reason why the next objective is to design a passive noise absorber in order to achieve more global attenuation in the high frequency range (*i.e.* above 1000Hz).

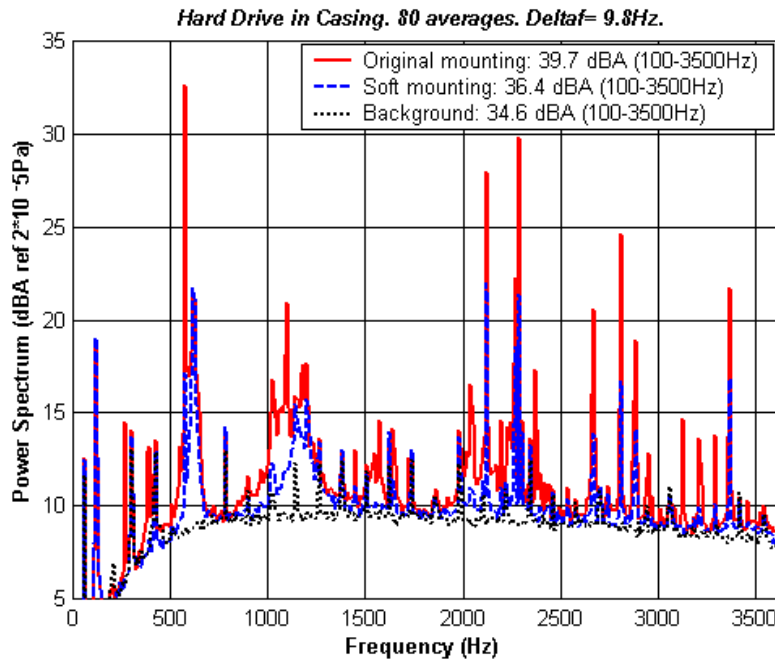


Figure 4.3 Hard drive structural decoupling effect on PC sound power level.

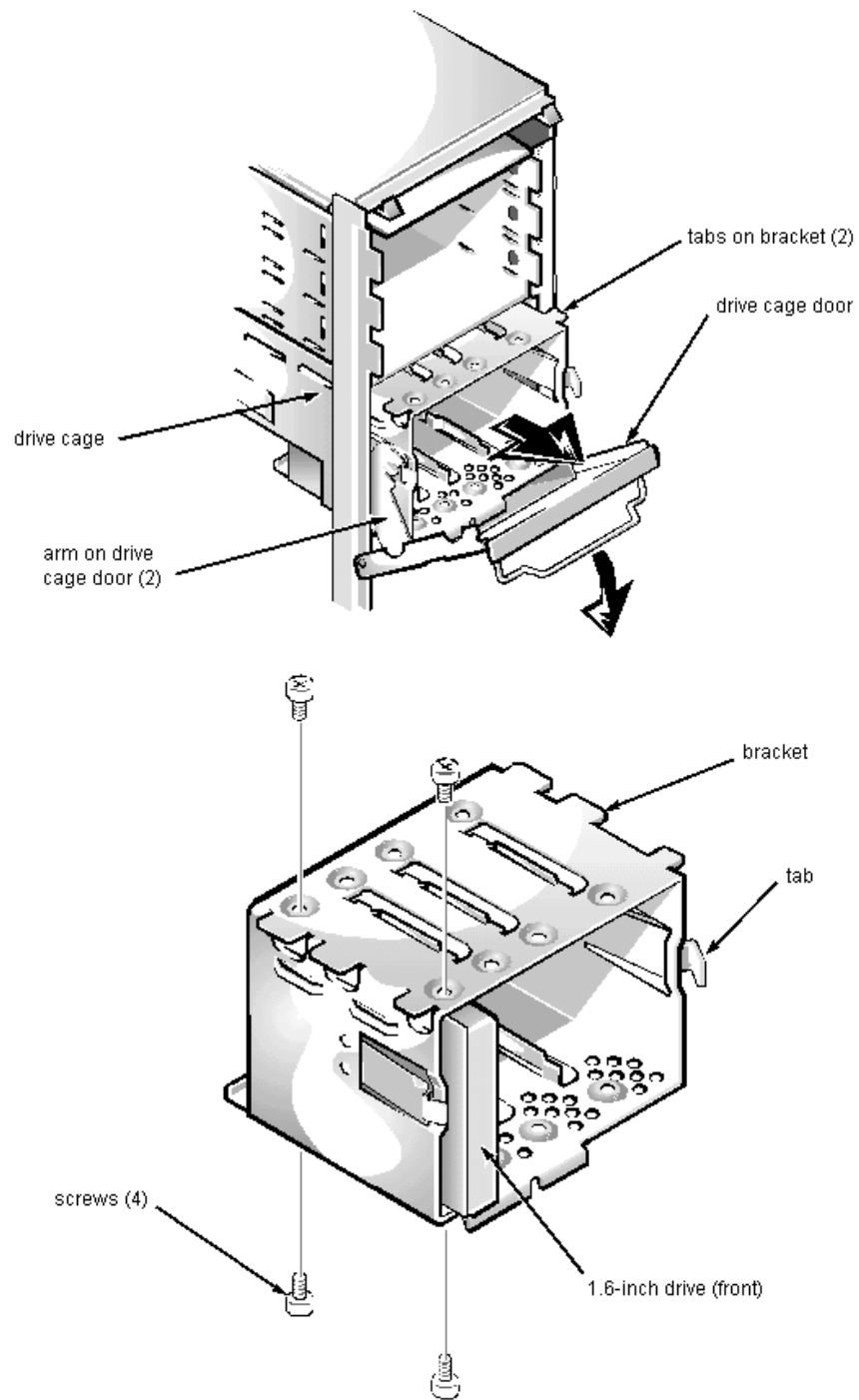


Figure 4.4 Hard drive mounting in standard bracket

4.2. Fan Ducting

4.2.1. Duct Acoustics

In order to design a lined duct that has good passive as well as active sound absorption characteristics, we should be aware of the way sound waves propagate in ducts and the techniques used to attenuate them. We start with an introduction to sound fields in rigid cavities to derive the notions of cut-on frequency, and propagative modes in ducts. We will then become familiar with dissipative silencers such as lined ducts and identify the key parameters in the design and performance. References [25] and [26] further develop duct acoustics and lined duct theory.

4.2.1.1. Theory of Wave Propagation in a Waveguide

a. Effects of Geometry

The Cartesian coordinates $\mathbf{x}$ as well as the associated particle velocity $\mathbf{u}$ are defined by:

$$\mathbf{x} = [x_1 \quad x_2 \quad x_3]^T \quad \mathbf{u} = [u_1 \quad u_2 \quad u_3]^T$$

We consider pressure fluctuations in a perfect gas undergoing adiabatic compression (no energy dissipated by the compressed element). In many books describing fundamental acoustics such as [26], it is shown that the three-dimensional wave equation in Cartesian coordinates can be expressed by:

$$\sum_{i=1}^3 \left(\frac{\partial^2 p}{\partial x_i^2} \right) - \frac{1}{c_0^2} \cdot \frac{\partial^2 p}{\partial t^2} = 0, \quad (4.1)$$

where $c_0 = \sqrt{\beta / \rho_0}$ is the phase speed for acoustic waves in fluids, related to the Bulk modulus and density.

It also shown in reference [26] that the pressure is related to the particle velocity by the linearized equation of momentum conservation, also called Euler's equation:

$$\rho_0 \cdot \frac{\partial u_i}{\partial t} + \frac{\partial p}{\partial x_i} = 0 \quad i = \{1 \ 2 \ 3\} \quad (4.2)$$

We consider a waveguide of constant cross-section and infinite length with perfectly rigid walls as shown in Figure 4.5. The cross-dimensions are L_1 in the x_1 direction and L_2 in the x_2 direction.

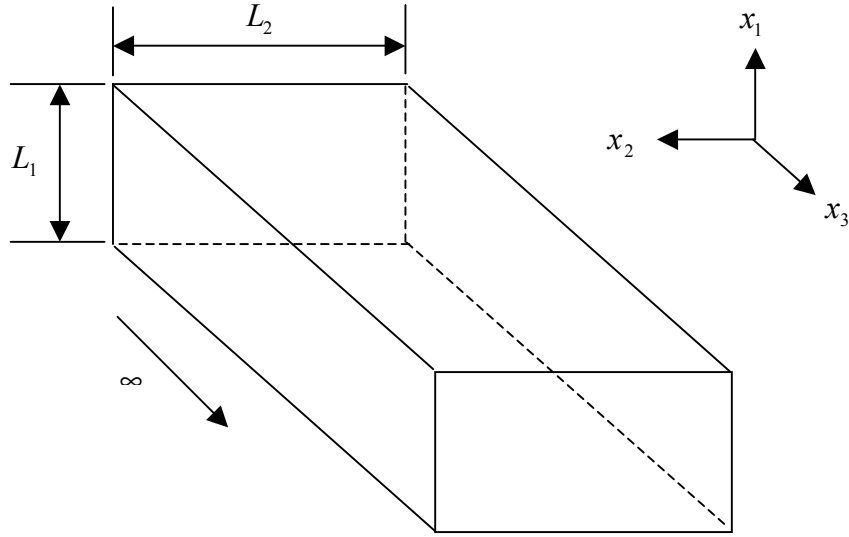


Figure 4.5 Waveguide modeling.

b. Cavity Modes

We consider harmonic waves:

$$p(\mathbf{x}, t) = p(\mathbf{x}) \cdot \exp(j \cdot \omega \cdot t) \quad (4.3)$$

$$\mathbf{u}(\mathbf{x}, t) = \mathbf{u}(\mathbf{x}) \cdot \exp(j \cdot \omega \cdot t) \quad (4.4)$$

By substituting the expression for the pressure above into the wave equation (4.1), it reduces to the Helmholtz equation:

$$\sum_{i=1}^3 \left(\frac{\partial^2 p(\mathbf{x})}{\partial x_i^2} \right) + k^2 \cdot p(\mathbf{x}) = 0 \quad (4.5)$$

Also, Euler's equation is now reduced to:

$$j \cdot \omega \cdot \rho_0 \cdot u_i(\mathbf{x}) + \frac{\partial p(\mathbf{x})}{\partial x_i} = 0 \quad i = \{1 \ 2 \ 3\} \quad (4.6)$$

More over, the complex pressure is assumed to be composed of three functions of a single variable:

$$p(\mathbf{x}, t) = \prod_{i=1}^3 (X_i(x_i)) \cdot \exp(j \cdot \omega \cdot t) \quad (4.7)$$

We can then derive from equation (4.5) a set of three differential equations:

$$\frac{\partial^2 X_i(x_i)}{\partial x_i^2} + k_i^2 \cdot X_i(x_i) = 0 \quad i = \{1 \ 2 \ 3\} \quad (4.8)$$

where the k_i are related by:

$$k = \sqrt{\sum_{i=1}^3 k_i^2} = \frac{\omega}{c_0} \quad (4.9)$$

Solutions to the set of differential equations (4.8) can be of the form:

$$X_i(x_i) = X_i \cdot \exp(-j \cdot k_i \cdot x_i) \quad i = \{1 \ 2 \ 3\} \quad (4.10)$$

$$X_i(x_i) = A_i \cdot \cos(k_i \cdot x_i) + B_i \cdot \sin(k_i \cdot x_i) \quad i = \{1 \ 2 \ 3\} \quad (4.11)$$

In order to identify A_i and B_i , we have to apply the boundary conditions. Since there is no boundary in the x_3 direction, standing waves will establish in the x_1 and x_2 directions, while a waveform will propagate in the x_3 direction.

When assuming that the walls are rigid, we impose the particle velocity to be zero at the boundaries. Then from equation (4.6):

$$\left(\frac{\partial p(\mathbf{x})}{\partial x_i} \right)_{x_i=0} = \left(\frac{\partial X_i(x_i)}{\partial x_i} \right)_{x_i=0} \cdot X_{j \neq i}(x_j) \cdot X_{k \neq i,j}(x_k) = 0 \quad i = \{1 \ 2\} \quad (4.12)$$

$$\left(\frac{\partial p(\mathbf{x})}{\partial x_i} \right)_{x_i=L_i} = \left(\frac{\partial X_i(x_i)}{\partial x_i} \right)_{x_i=L_i} \cdot X_{j \neq i}(x_j) \cdot X_{k \neq i,j}(x_k) = 0 \quad i = \{1 \ 2\} \quad (4.13)$$

By substituting the expression (4.11) into (4.12) and (4.13), we deduce that $X_1(x_1)$ and $X_2(x_2)$ reduce to cosine functions with specific wavenumbers k_{1l} and k_{2m} which permits to express (4.7) differently:

$$p_{lm}(\mathbf{x}, t) = A_{lm} \cdot \cos(k_{1l} \cdot x_1) \cdot \cos(k_{2m} \cdot x_2) \cdot \exp(j \cdot \omega \cdot t - j \cdot k_3 \cdot x_3), \quad (4.14)$$

where:

$$k_{1l} = \frac{l \cdot \pi}{L_1} \quad l = \{0 \ 1 \ 2 \ \dots\} \quad (4.15)$$

$$k_{2m} = \frac{m \cdot \pi}{L_2} \quad m = \{0 \ 1 \ 2 \ \dots\} \quad (4.16)$$

We now define a transverse wavenumber k_{lm} :

$$k_{lm} = \sqrt{k_{1l}^2 + k_{2m}^2} \quad (4.17)$$

The wavenumber in the direction of propagation k_3 is related to the transverse wavenumber k_{lm} by the expression in equation (4.9). Thus:

$$k_3 = \sqrt{\left(\frac{\omega}{c_0} \right)^2 - k_{lm}^2} \quad (4.18)$$

c. Cut-On Frequencies

It can be proved that any complex wave in the duct can be expressed as a sum of modes:

$$p(\mathbf{x}, t) = \sum_{l,m} p_{lm}(\mathbf{x}, t) = \sum_{l,m} A_{lm}(x_1, x_2) \cdot \exp(j\omega t - jk_3 x_3), \quad (4.19)$$

where:

$$A_{lm}(x_1, x_2) = \cos(k_{1l} \cdot x_1) \cdot \cos(k_{2m} \cdot x_2) \quad (4.20)$$

In function of the sign of the argument in the square root of equation (4.18), the wavenumber in the direction of propagation k_3 can be negative or positive. We can thus now prove that depending on the frequency, any mode composing the pressure field will be propagating or evanescent.

Let first consider a mode (l, m) such as $\omega/c_0 > k_{lm}$ which implies that k_3 is positive and purely real. In such case, the mode (l, m) is propagating in the k_3 direction.

On the other hand, assume that the $\omega/c_0 < k_{lm}$. Then equation (4.18) becomes:

$$k_3 = \sqrt{\left(\omega/c_0\right)^2 - k_{lm}^2} = \sqrt{j^2 \cdot \left(k_{lm}^2 - \left(\omega/c_0\right)^2\right)} = \pm j \cdot \sqrt{\left(k_{lm}^2 - \left(\omega/c_0\right)^2\right)} \quad (4.21)$$

In that case, k_3 can be positive or negative and is purely imaginary. The wave expression $p_{lm}(\mathbf{x}, t)$ for the mode (l, m) then becomes:

$$p_{lm}^+(\mathbf{x}, t) = A_{ml}(x_1, x_2) \exp\left(-\sqrt{k_{lm}^2 - \left(\omega/c_0\right)^2} x_3\right) \exp(j\omega t) \quad (4.22)$$

$$p_{lm}^-(\mathbf{x}, t) = A_{ml}(x_1, x_2) \exp\left(-\sqrt{k_{lm}^2 - \left(\omega/c_0\right)^2} x_3\right) \exp(j\omega t) \quad (4.23)$$

The wave expression $p_{lm}^+(\mathbf{x}, t)$ physically corresponds to a standing wave in the (x_1, x_2) plane, propagating in the positive x_3 direction but with a modal amplitude decaying exponentially. The wave expression $p_{lm}^-(\mathbf{x}, t)$ corresponds to a similar evanescent wave form but that is propagating in the negative x_3 direction. The mode (l, m) is here called evanescent since no power is propagating in the waveguide.

We can thus define the cut-on frequency ω_{lm} for the specific mode (l, m) as the frequency above which it will constitute a propagative wave in the guide. At the cut-on frequency, the mode (l, m) is a pure standing wave in the (x_1, x_2) plane since:

$$k_3 = \sqrt{\left(\frac{\omega_{lm}}{c_0}\right)^2 - k_{lm}^2} = 0 \quad (4.24)$$

which implies that:

$$p_{lm}(\mathbf{x}, t) = A_{lm} \cdot \cos(k_{1l} \cdot x_1) \cdot \cos(k_{2m} \cdot x_2) \cdot \exp(j \cdot \omega \cdot t) \quad (4.25)$$

We express the cut-on frequency in function of the mode order:

$$f_{lm} = \frac{\omega_{lm}}{2\pi} = \frac{c_0 \cdot k_{lm}}{2\pi} = \frac{c_0}{2} \sqrt{\left(\frac{l}{L_1}\right)^2 + \left(\frac{m}{L_2}\right)^2} \quad (4.26)$$

When $l = m = 0$, $k_{1l} = k_{2m} = 0$. Thus, the pressure field is a simple plane wave since the amplitude is constant in the (x_1, x_2) plane:

$$p_{lm}(\mathbf{x}, t) = A_{lm} \cos(0 \cdot x_1) \cos(0 \cdot x_2) \exp(j\omega t - jk_3 x_3) = A_{lm} \cdot \exp\left(j\omega\left(t - \frac{x_3}{c_0}\right)\right) \quad (4.27)$$

Such a wave has no cut-on frequency. Plane waves are propagating in the waveguide at all frequencies. By selecting appropriate cross-dimensions for the waveguide, we can ensure that only plane waves will be propagating in the duct below a chosen frequency. From equation (4.26), we can deduce the frequency above which the first-cross mode ($l = 1$ or $m = 1$) will propagate in the duct:

$$f_c = \frac{c_0}{2 \cdot \max(L_1, L_2)} \quad (4.28)$$

Equation (4.28) constitutes an important design parameter for us, considering that a simple Active Noise Control system will be integrated to the duct. Indeed, it will be explained in Chapter 5 that the objective of the Active Noise Controller in the duct is to generate a waveform in the duct what will cancel out the propagating disturbance. If the acoustic problem is limited to plane waves propagating in the duct, then the use of a single secondary source (such as a loudspeaker operating in its piston range) generating plane waves will be sufficient to control the noise. Also, since the modal amplitude will be constant in the duct cross-section, a single reference and single error microphones will be needed in the duct in order to obtain sufficient information about the propagating disturbance and the control performance.

On the other hand, if cross-modes are propagating in the duct in the frequency range of the applied active noise control, then a single secondary noise source will not be sufficient to reproduce the complex disturbance field. Also, multiple sensors will be required to obtain a complete description of the disturbance field since its amplitude and phase will not be constant in any of the transverse plane.

Our objective is to implement in the duct a simple Active Noise Controller with a single reference sensor, error sensor and secondary source (c.f. Chapter 5). Thus, an important goal in the duct design will be to ensure that only plane waves are propagating in the waveguide in the frequency range of Active Noise Control (typically up to 1000-1500Hz).

4.2.1.2. Dissipative Silencers

The following discussion is based on the book “Noise and Vibration Control Engineering” from Beranek and Ver [25]. Two main types of silencers can be distinguished for the control of noise propagating in ducts. Reactive type silencers, such as expansion chamber mufflers do not dissipate energy but rather reflect sound waves toward the noise source, upstream. On the other hand, there exists dissipative silencers such as lined ducts, parallel baffle silencers or circular ducts as shown in Figure 4.6. Such systems attenuate sound due to friction losses between the air particles and the fibrous materials constituting the liner applied usually to the duct walls or center body.

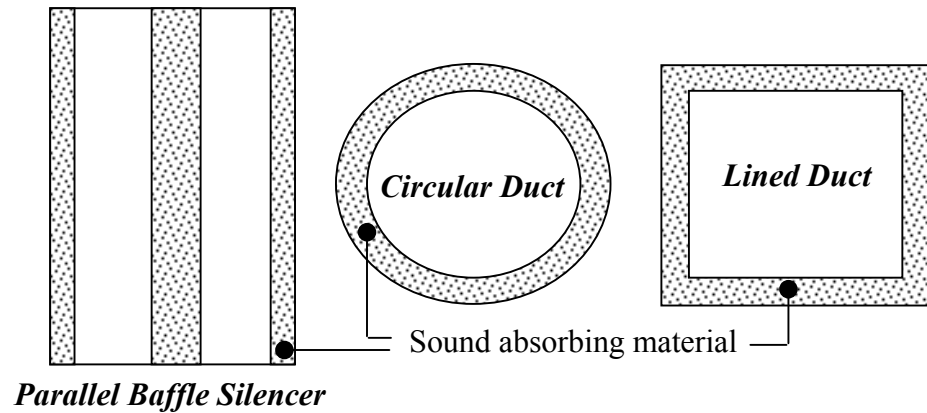


Figure 4.6 Different designs of dissipative silencers.

In this project, lined ducts will be used. We will first define technical terms used to characterize the silencer properties and efficiency. The objective here is not to perfectly predict the attenuation provided by the lined duct but rather identify the important parameters in the duct geometry that influence its sound absorbing characteristics. Thus, based on reference [25], we will review some practical requirements to achieve certain sound attenuation goals in parallel-baffle type silencers. Those requirements can be easily transposed to circular or, in our particular case, lined ducts.

a. Key Performance Parameters

Insertion Loss

The insertion loss of a silencer connected to a duct (in dB) is defined [25]:

$$IL = 10 \cdot \log_{10} \left(\frac{W_o}{W_M} \right), \quad (4.29)$$

where W_M and W_o represent the sound power radiated from the duct with and without the silencer respectively. One other expression is given by:

$$IL = -10 \cdot \log_{10} \left(\frac{W_{SG}}{W_0} + 10^{-(\Delta L_l + \Delta L_{ENT} + \Delta L_{EX})/10} \right), \quad (4.30)$$

where W_{SG} is the sound power generated by the flow existing in the silencer, ΔL_l is the attenuation of the silencer of length l , ΔL_{ENT} and ΔL_{EX} are the entrance and exit losses. Since in our fan noise application, the flow velocity in the silencer will be fairly low, flow noise can be neglected. Then, equation (4.30) simplifies to:

$$IL \approx \Delta L_l + \Delta L_{ENT} + \Delta L_{EX} \quad (4.31)$$

Entrance and Exit Losses

Usually, entrance losses ΔL_{ENT} are neglected when the incident sound energy in the silencer is in the form of a plane wave normally incident on the silencer entrance. However, in our particular problem, the lined ducts will be used to attenuate noise radiated by fans mounted on the PC casing and thus originally radiating in a half-space free field. The original sound field, constituted by spherical waves will be transformed into a plane-wave sound field (plus eventually higher order cross-modes propagating down the duct above certain cut-off frequencies). The change in the sound field will have effects on the entrance loss. For example, the conversion of a semi diffuse sound field (when many high order modes are excited in a cavity) in the entrance duct into a plane-wave field in the duct typically results in an entrance loss of 3 to 6dB [25].

The exit losses ΔL_{EX} are important when the dimensions of the open end of the lined duct are small compared with the wavelength. The exit loss is then predominantly determined by the end reflections.

The relative importance of the entrance and exit losses diminishes as the silencer length increases because, contrary to ΔL_l , those losses are independent of the silencer length.

Silencer Attenuation

The silencer attenuation ΔL_l (in dB) is proportional to its length l and to the lined perimeter of the passage P . On the other hand, it is inversely proportional to the cross-sectional area of the passage A :

$$\Delta L_l = \frac{P}{A} \cdot l \cdot L_h \quad (4.32)$$

The parameter L_h , referred as the attenuation per channel height, depends on the geometry of the passage, on the acoustical properties of the liner used, on the frequency and the temperature and velocity of the flow in the passage. In the next section, we will evaluate this key performance parameter in the case of parallel baffle silencers.

Pressure Drop

The total pressure drop (in Pa) across a muffler is due to the entrance, exit and friction losses:

$$\Delta p_T = \frac{1}{2} \cdot \rho \cdot v_p^2 \cdot \left(K_{ENT} + K_{EX} + \frac{P}{A} \cdot l \cdot K_F \right), \quad (4.33)$$

where ρ is the density of the gas, and v_p its face velocity in the passage of the silencer.

The constants K_{ENT} and K_{EX} are the entrance and exit head loss coefficients, depending on the geometry of the passage configuration. K_F represents the frictions losses.

We observe, that both the pressure drop due to friction losses and silencer attenuation are proportional to $\frac{P}{A} \cdot l$. Thus, if the goal is to maximize the attenuation without inducing too much pressure drop in the silencer, we should try to maximize the attenuation per channel height rather than the ratio $\frac{P}{A} \cdot l$.

b. Attenuation Performance for Parallel-Baffle Type Silencers

The parallel-baffle silencer as shown in Figure 4.7 is very frequently used because of its good acoustical performance and low cost. Such a type of dissipative silencer will not be used in practice, but rather a lined duct. Nevertheless, it is useful to learn about the design rules for this type of silencer since they will equally apply to the lined duct.

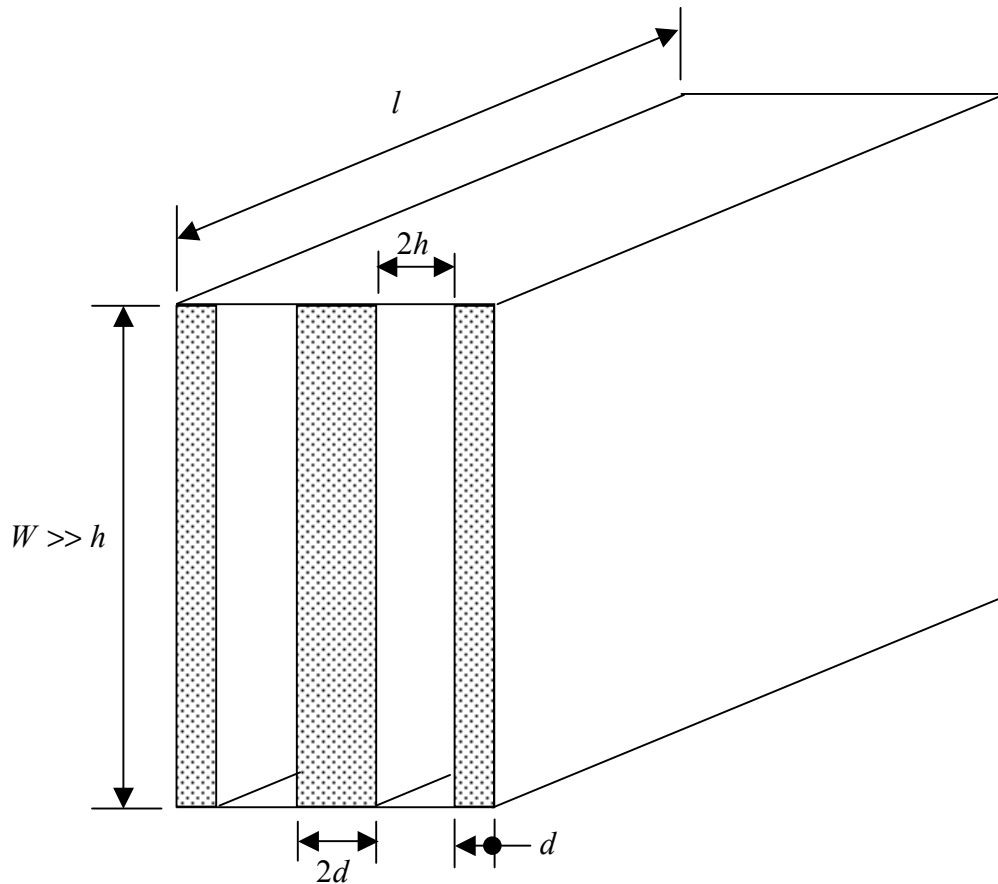


Figure 4.7 Parallel-baffle type silencer

It is suggested [25] that the sound energy traveling in the passages of the parallel-baffle silencer can be effectively attenuated over a broad range of frequencies if the sound enters the porous sound absorbing material and if a substantial part of energy of the incident wave is dissipated in the absorbing layer before it reenters the passage.

In order to have some energy penetrate in the sound absorbing layer, the passage height should be small compared to the wavelength ($2 \cdot h < \lambda$). Also, the porous sound absorbing material should be open enough so that the sound wave enters it rather than be reflected at the interface due to a too high impedance break between the mediums. The material should thus be soft and present low flow resistivity.

In order to have some energy absorbed by the porous layer in which the wave is traveling, this material should present a moderate flow resistivity. Thus, the requirements of easy sound penetration and high energy dissipation are contradictory. The choice of baffle thickness and flow resistivity of the porous material is always a compromise.

In function of the shape of the silencer geometry, the attenuation per channel height L_h can be maximized over a certain frequency range. To provide reasonable attenuation at lower frequencies, it is recommended [25] that the baffle thickness be of the order of $\frac{1}{8}$ of the wavelength. The relationship is given by equation (4.34). For example, to passively attenuate sound waves whose frequency is around 1000Hz, the baffle should be theoretically at least 43mm thick:

$$d > \frac{c}{8 \cdot f_{low}} \approx \frac{340}{8 \cdot 1000} = .0425 \quad (4.34)$$

On the other hand, in order to provide reasonable attenuation at higher frequencies, the passage height h should be smaller than the wavelength as shown in equation (4.35). Assuming that the duct should be effective up to 5kHz, the passage height needs to be less than 70mm:

$$h < \frac{c}{f_{high}} \approx \frac{340}{5000} = 0.068 \quad (4.35)$$

We define the normalized flow resistance R :

$$R = \frac{R_l \cdot d}{\rho \cdot c}, \quad (4.36)$$

where R_l is the flow resistivity of the sound absorbing material, typically between 5000 and 12000 $N \cdot s/m^3$ ($0.3\rho \cdot c/in$ to $0.75\rho \cdot c/in$). To allow proper penetration of the sound in the absorbing layer, it is recommended [25] that the normalized flow resistance be comprised between 2 and 6, which can be translated in terms of baffle thickness:

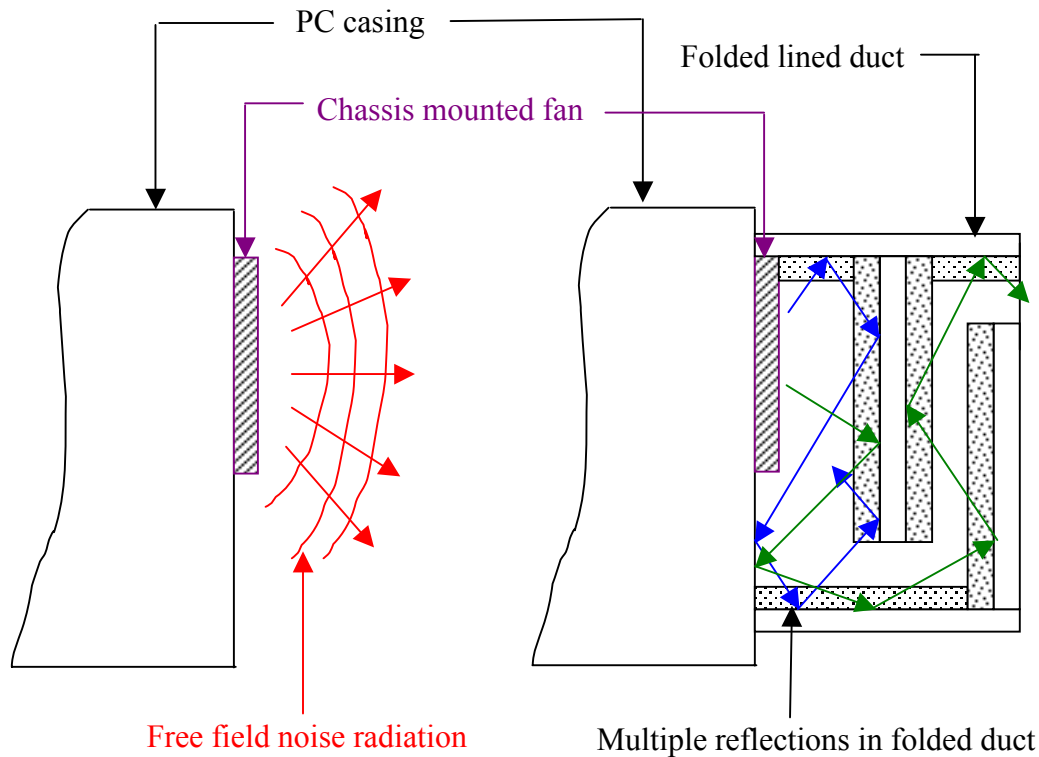
$$2 \cdot \frac{\rho \cdot c}{R_l} \leq d \leq 6 \cdot \frac{\rho \cdot c}{R_l} \quad (4.37)$$

Overall, the attenuation vs. frequency curves for parallel silencers have the shape of a bandpass-filter with a maximal efficiency at frequencies where the wavelength is between 2 and 4 times the passage height ($2 \cdot h < \lambda < 4 \cdot h$). The larger the flow resistance R , the higher the central frequency at which the silencer will be the most effective. Also the range of frequencies that will be well attenuated gets broader when the ratio d/h is increased. Typically, $0.5 < d/h < 2$ to ensure good efficiency.

For our problem, we cannot afford to optimize the previous parameters for the passive attenuation of low frequencies (below 1000Hz) since the duct would then become too large and could not be integrated in the PC. Moreover, our silencer will be added to the original fan cooling system. We need to design a silencer that will not penalize the thermal characteristics of the cooling system. For those reasons, our practical design will rather be a lined duct with a thin sound absorbing layer and large passage section in order to minimize the pressure drop in the silencer. Its attenuation properties will rather be due

to entrance and exit losses as well as reflections resultant to multiple folds rather than pure energy dissipation in the liner as can be seen in Figure 4.8.

Figure 4.8 Illustration of multiple reflections phenomenon in a folded duct.



Our main goal in the duct design will be to ensure that only plane waves are propagating in the frequency range of active noise control which will constrain the allowable dimensions for the ducts. As we have seen in the sources identification section, some of the noise sources radiate significant amounts of energy below 1000Hz, that cannot be controlled passively. Active Noise Control techniques are specially suited for this kind of application where we want to achieve good low frequency noise control using little space.

4.2.2. First Duct Prototype

4.2.2.1. Constraints

One of the major constraints in the integration of ducts in a commercial PC is flexibility. The machine we have been working on is a mini-tower. Due, to its small external dimensions (220 by 440 by 440 mm), no much space is available inside the casing for adding a duct. Moreover, any available space inside the casing is reserved for expansion boards, extra storage devices, etc. In order not to sacrifice those expansion slots, the ducts cannot be integrated inside the current casing. This leads to a size constraint; the ducts of a fairly small size are required, so that the added volume to the PC is not too large.

The other major constraint is the cooling issue. The addition of ducts to the PC air cooling circuit will eventually reduce the airflow performance. For example, if the duct cross-section area is much smaller than that of the fan, its use will lead to a reduction in the airflow since the fan speed is kept unchanged. Thus, a certain duct internal cross-section needs to be maintained. Considering fixed external cross dimensions, this translates into limited acoustic performance. Indeed, to increase the attenuation in the lined duct, we need to maximize the ratio P/A and thus minimize the duct cross-section as can be seen in Figure 4.9. Also, the liner should be as thick as possible in order to be efficient at lower frequencies. In theory, for normal incidence wave the acoustical foam thickness should be at least a quarter of a wavelength in order to be effective. In practice, the incident waves will have some angle of incidence, and thus be attenuated by the liner, even at low frequencies.

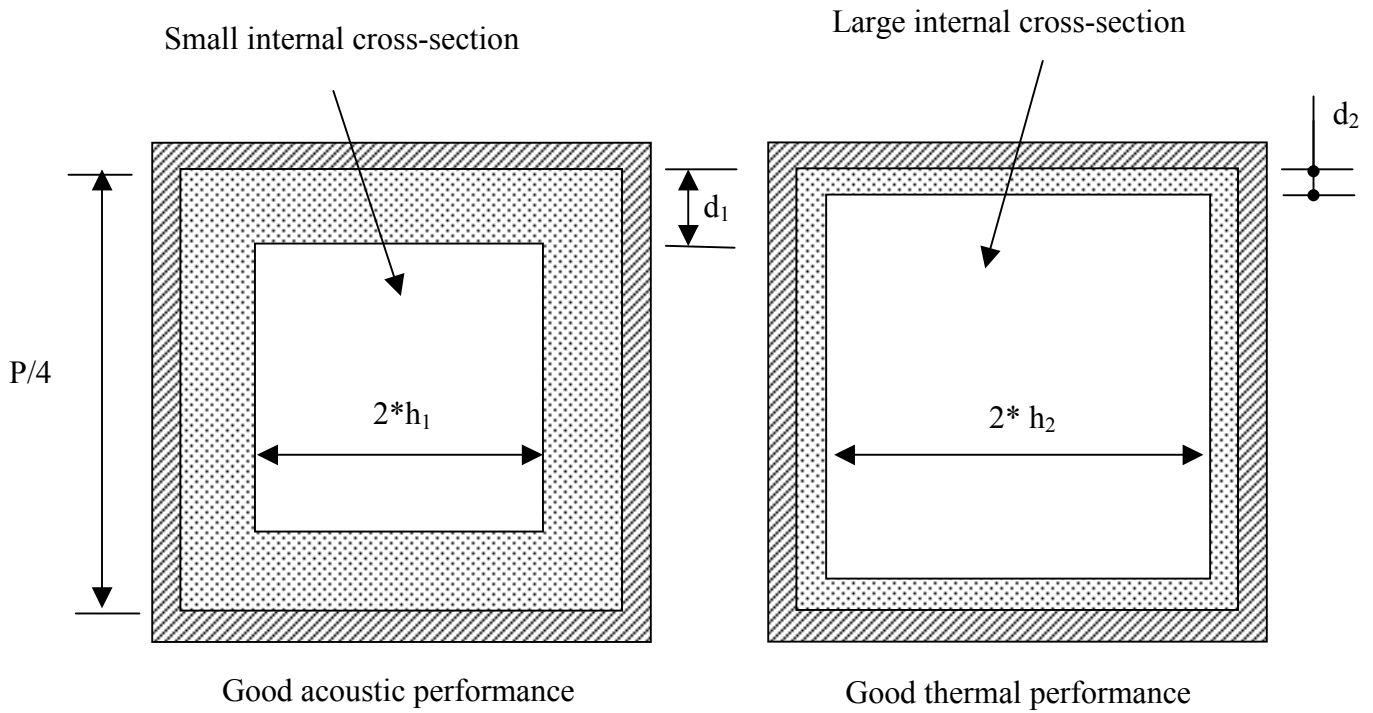


Figure 4.9 Compromises between thermal and acoustic performance in lined ducts.

4.2.2.2. Geometry

The first lined duct prototype was not designed to be integrated inside the PC casing. It is a folded lined duct attached to the rear panel of the PC casing, covering both the power supply and processor fans as can be seen in Figure 4.10, Figure 4.11, and Figure 4.12.

We tried to maximize the ratio of liner thickness d to duct internal cross-dimension h so that significant attenuation is obtained over a wide range of frequencies. The ratio d/h defined previously is between .25 and 1 as suggested for parallel baffle silencer. The maximum attenuation should be achieved for wavelengths smaller than four times $h_{\max}$, which corresponds to frequencies above about 2000Hz ($h_{\max} = 47.5\text{mm}$). Due to size constraints, we could not maintain an internal duct cross-section as large as the fans cross-section. This might be a limiting factor for the volume of airflow exiting the PC casing and will be discussed below.

A 3" diameter electro-dynamic driver (Radio-Shack 40-254) in sealed enclosure is integrated in the duct for Active Noise Control purposes. The duct needs to be as long as possible so that the ANC system can operate effectively for broadband noise control as will be seen in Chapter 5. Also, the overall silencer attenuation is proportional to its length. The lined duct is folded so that it can be practically integrated to the PC rear panel without obstructing the expansion board area or any motherboard connector. The other advantage of multiples folds is to provide some extra attenuation due to reflections back to the source as seen earlier. The lined duct is thus not only a dissipative but also a reactive silencer.

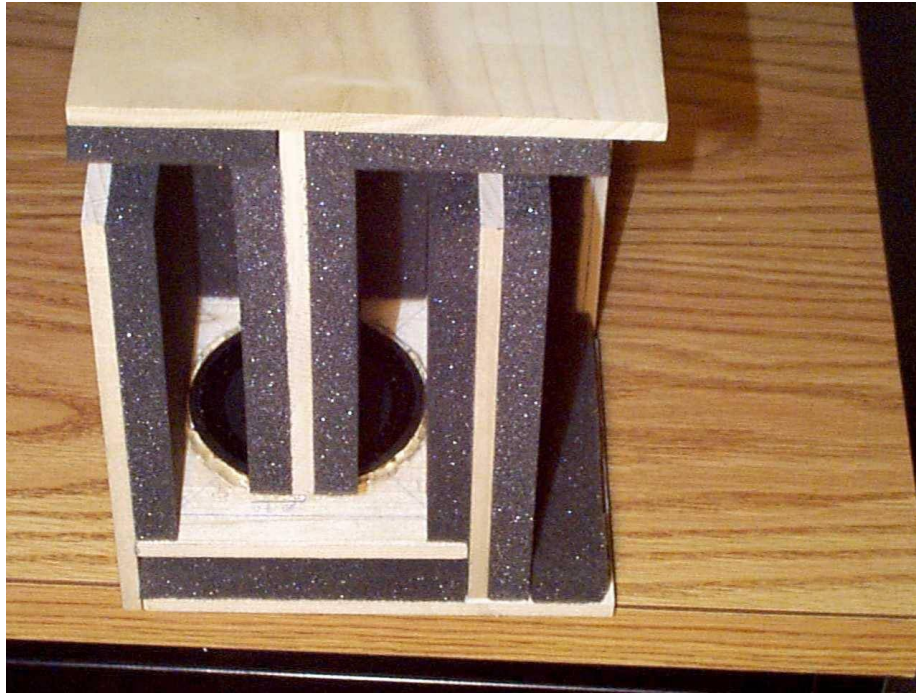


Figure 4.10 Picture of the first outlet duct prototype.

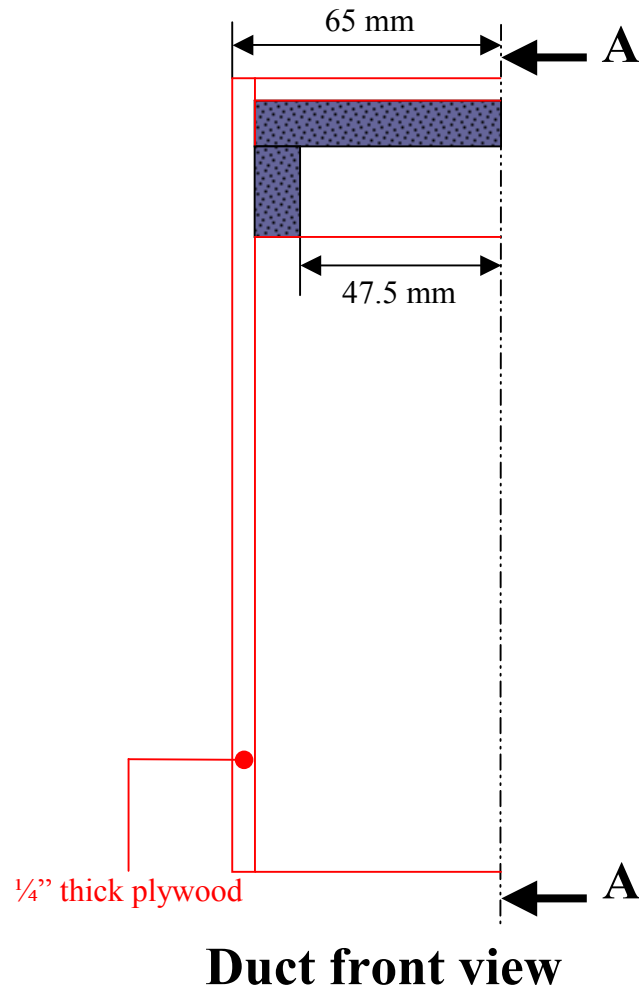


Figure 4.11 Half front view the first outlet duct prototype.

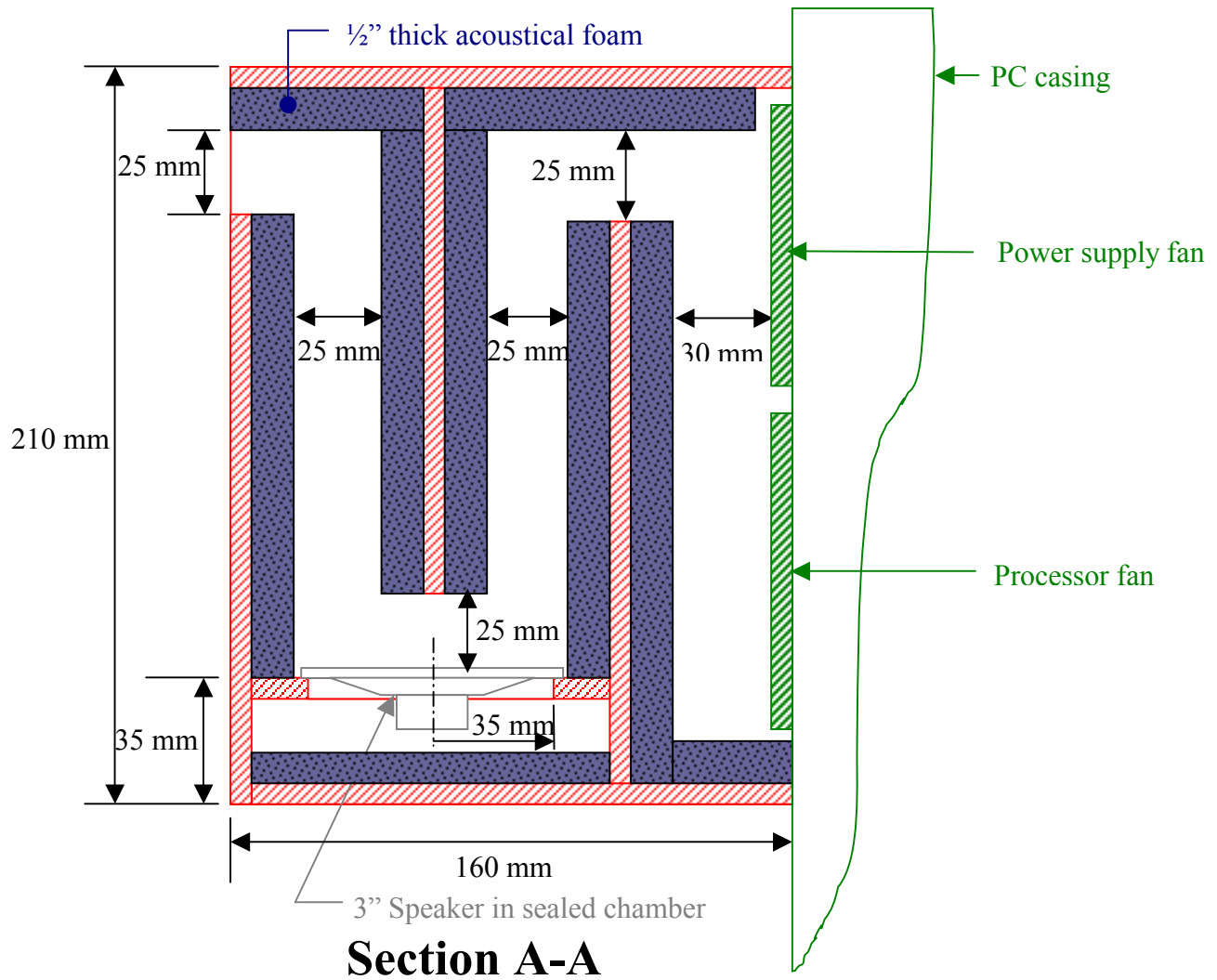


Figure 4.12 Section cut of the first outlet duct prototype.

4.2.2.3. Acoustical performance

a. Pseudo-Insertion Loss

For the first duct prototype, the true insertion loss was not measured. As we have seen before, the insertion loss of a silencer is evaluated by measuring the attenuation in the sound power level radiated by the wave-guide due to the addition of the silencer. In the present case, the situation is quite different since the sources are baffled in their original configuration (fans are located on the PC rear panel). We could evaluate a quantity similar to the insertion loss by measuring the sound power level radiated by the sources in the PC with and without the duct. Unfortunately, such a method will correctly characterize the duct only if the major radiation path is the free field noise radiation on the PC rear panel, which is not the case in practice. For the following results, we will simply measure the variation in the sound pressure level at different locations due to the addition of the lined duct as shown in Figure 4.13.

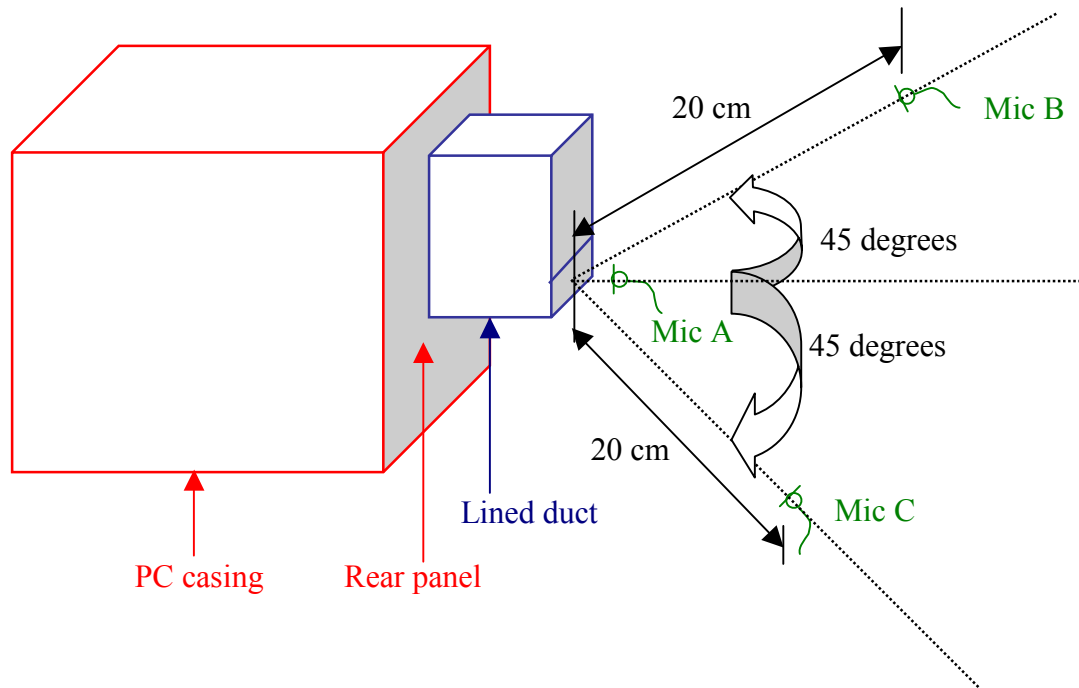


Figure 4.13 Pseudo insertion loss measurement setup.

In order to evaluate the duct performance over a wide frequency range, the actual noise sources are replaced by a speaker. We measured the effect of the duct on the sound pressure level at the rear of the PC when a speaker, with its sealed enclosure, is mounted at the processor fan location in the PC casing. The noise source is a 4" diameter driver. Contrary to the fan that has bipolar radiation characteristics, the speaker is mounted in a sealed enclosure, thus only radiating in a half-space free field when it is baffled. The sound pressure level is measured at the three locations shown Figure 4.13 when the noise source is fed with white noise. The signals are recorded at a sample rate of 20kHz, which enables a measurement of the insertion loss over a wide frequency range. The pseudo insertion loss computed, corresponds to the difference in SPL level when the duct is attached or not as shown in Figure 4.14.

In the far field (locations b and c), 20 to 40dB of attenuation is achieved above 1200Hz, which proves that the silencer is effective. Nevertheless, at the duct outlet (location a), we observe an amplification of the sound pressure level below 1000Hz. One possible explanation of this phenomenon is that the loudspeaker radiation impedance is increased at low frequencies when replacing the plane baffle by the duct. Then, for a given volume velocity of the speaker cone, a higher sound pressure level is generated. However, since this observation was made only at one microphone location, in the near field of the sound source, this increase in near field sound pressure might not lead to increased transmitted power. For example, the radiation impedance might be higher when a duct is present, but purely reactive, and in such case, the effective acoustic energy flow outside the duct is null since the complex pressure and velocity are in quadrature [26]. This would explain why the insertion loss is "positive" at all frequencies at the two measured far field positions.

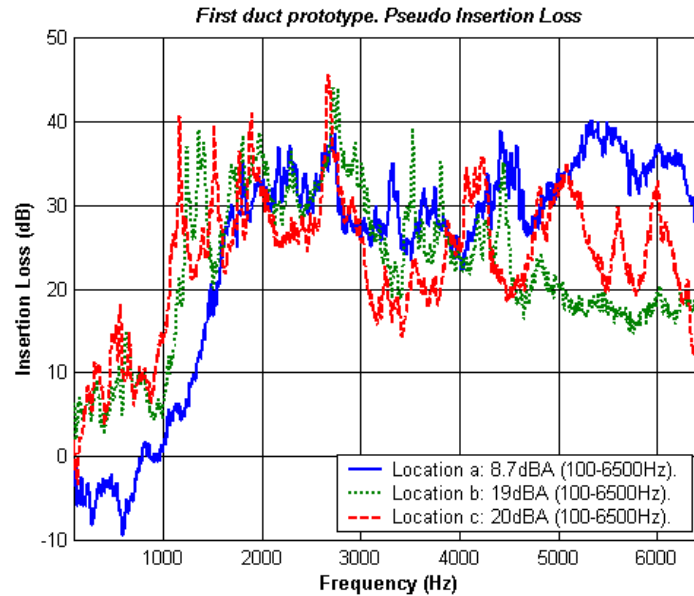


Figure 4.14 Pseudo insertion loss results.

b. Performance limitations

The first problem with this duct prototype is that its internal cross-section is much smaller than that of the fans it is covering. Indeed the fans area is 50 (power supply fan) + 65 (processor fan) = 115 cm² where as the internal duct cross-section is 22.5 cm². We believe this does affect the performance of the air cooling system. We only monitored the temperature when the processor was not in heavy use. The temperature remained between 85 and 90°F inside the PC casing when the duct was added (the ambient temperature was 75°F). The temperature in the duct was 95°F in the upstream section and 90°F at the outlet. Thus, the temperature did not significantly change when the duct was added. Even though the variations in volume of airflow evacuated from the PC casing were not measured, we suspect that it is affected by the duct presence.

The second problem is that the 3" diameter speaker integrated to the duct has a low radiation efficiency at low frequencies as well as limited power handling capacities. Such a weak transducer will lead to control authority problems in the Active Control of Noise for frequencies down to 200Hz. The control filter will compensate for the limited low frequency output of the driver to some extent but we will observe that, either the

controller, or the transducer itself, saturate at low frequencies. In the case of the controller, saturation occurs when output signal requirements exceed the maximal output voltage. In the case of the loudspeaker, saturation occurs if the cone displacement requirements are larger than the loudspeaker can handle. Typically, the smaller the cone, the bigger the saturation problems due to the limited volume velocity generated. In our case, the driver was chosen for its small size and weight upon testing but will reveal unsuitable for our control purposes.

The last problem is that the duct is only controlling the noise radiated by the fans at the air outlet location. We know that other radiation paths such as the air inlet in the PC front panel are also contributing to the overall noise radiated. Thus the previous limitations lead us to the design of new duct prototypes in an attempt to resolve those problems.

4.2.3. Second Duct Prototype

4.2.3.1. Geometry

The first major modification for the new prototype is that we built not only an outlet but also an inlet duct. The outlet duct location remained unchanged, while the inlet duct was integrated to the front bezel of the PC casing based on guidelines provided by Intel as can be seen in Figure 4.17. We had to relocate the hard drive since the inlet duct would occupy the space originally reserved for the internal drive bay. The processor fan was also relocated and is now soft-mounted on the inlet duct as shown in Figure 4.18. The air cooling configuration could be called push-pull since the processor fan is now blowing air into the PC casing where as the power supply fan is still evacuating the heat out of the PC casing. We suppressed any other air inlet (such as on the base of the PC casing) since the objective is to enable cooling exclusively through the new ducts. Provided that the PC casing presents a large transmission loss, the only way airborne noise can be radiated from the PC is then through the lined ducts.

The dimensions of the ducts are shown in Figure 4.19 through Figure 4.22. In terms of airflow, we believe the new ducts perform much better than the previous prototype. The design was based on the guidelines provided by Intel regarding fan ducting. In order to achieve good air cooling performance, the duct should have an internal cross-section of similar size to the fan it is covering. The outlet duct has an internal cross-section of 45 cm^2 . The power supply block fan is the only source of airflow in this duct with an area of 50 cm^2 which is fairly close. The inlet duct has a minimal internal cross-section of 56 cm^2 which is to be compared to the processor fan area of 65 cm^2 . Unfortunately, the volume of airflow in the duct was not measured since no typical performance target that would ensure adequate cooling of the PC was known. At least, we expect the ducts to not have a significant impact on the fans performance. The number of folds is also an other parameter that can affect the cooling performance. While the new outlet duct prototype only contains one fold, the inlet duct has a L shape. Thus, we believe such a limited number of folds should not significantly affect the cooling performance via reduced flow restriction.

In terms of passive noise control performance, the new ducts should not be performing as well as the old one. The first reason is that the liner used is twice as thin ($\frac{1}{4}''$ versus $\frac{1}{2}''$) where as the internal duct cross-section has been significantly increased. For the first duct prototype, we would quantify the ratio of liner thickness d to duct internal cross-dimension h . The new ducts (especially the inlet duct) have a more complicated design, very different from typical parallel baffle silencers. Thus the proportion of liner is now quantified in terms of percentage of the total duct cross-section. An example is shown in Figure 4.15. Where as the first duct prototype was 40% open, which ensured a good passive performance over the widest range of frequencies [25], the new outlet and inlet ducts are approximately 70% open which will eventually make them perform well in a narrower frequency range. We will try to quantify the ducts performance in the next section.

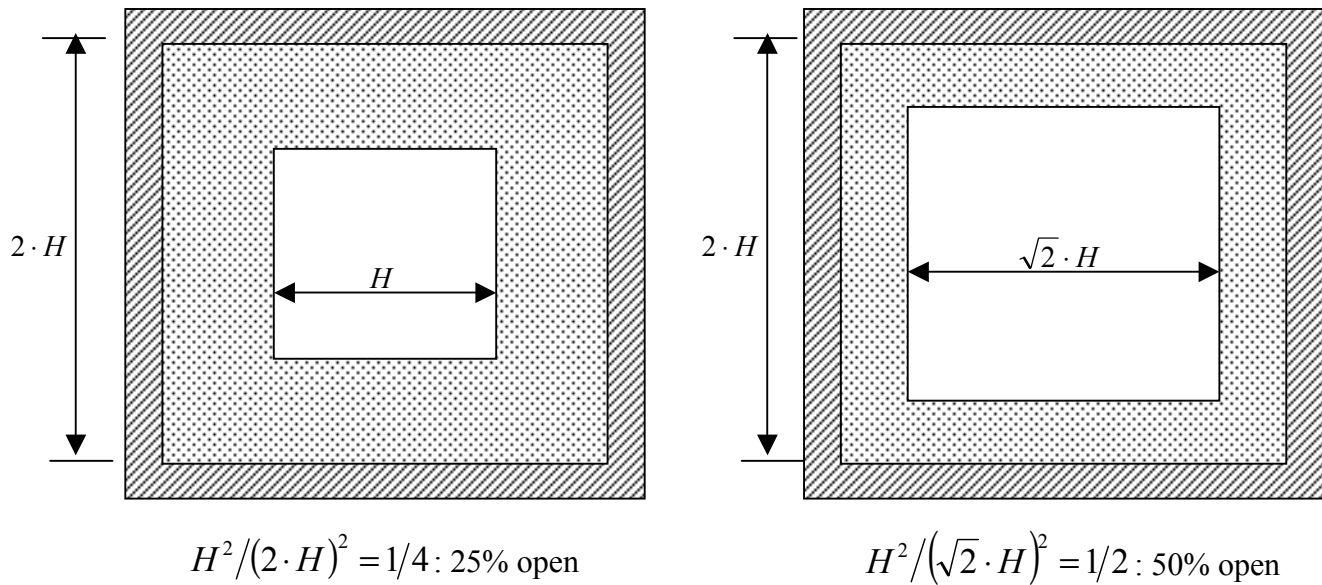


Figure 4.15 Illustration of the calculation of liner percentage in a duct.

The last major difference between the new and old duct prototype is the electro-dynamic driver used for ANC purposes. The new speaker (Radio-Shack Ref#40-1197) is larger in size, uses a stronger magnet and larger diaphragm as can be seen in Figure 4.16. It has much better performance characteristics as can be seen in Table 4.1 (manufacturer data).

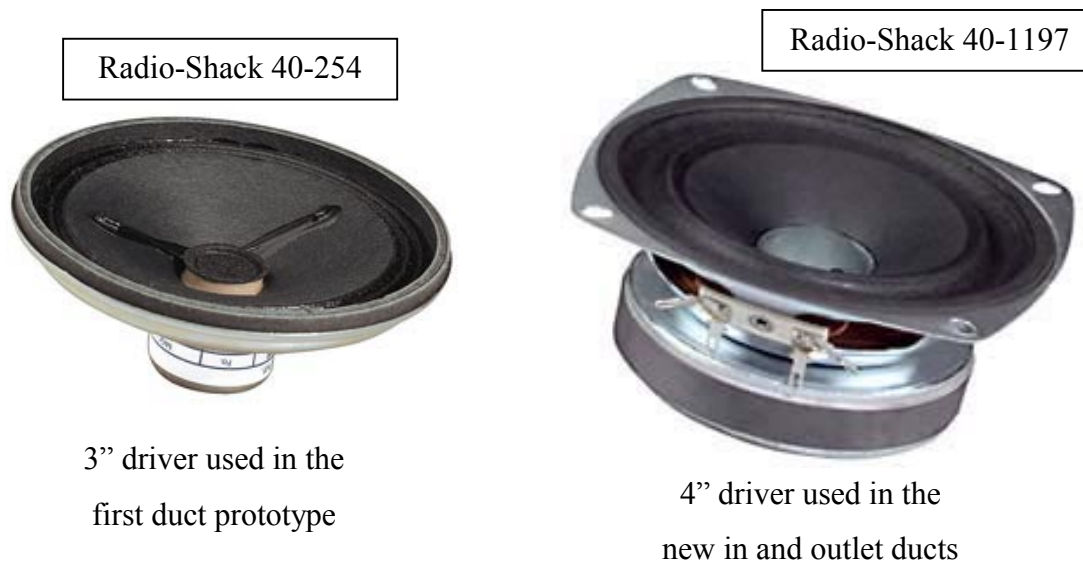


Figure 4.16 Picture comparing the old and new speakers used in the ducts.

Radio Shack reference:	RS 40-257	RS 40-1197
Speaker diameter:	3 inches	4 inches
Magnet weight:	10g	190g
Total weight:	<100g	590g
Power Handling (continuous)	2W	3W
Peak power:	4W	15W
Sensibility	Not available	90dB/W (at 1 meter)
Frequency range:	230-8000Hz	80-15000Hz
Retail price:	\$5	\$12

Table 4.1 Speakers performance comparison table

As can be seen in Table 4.1, the increase in speaker performance is achieved at the expense of weight, size, and cost. However, it will be shown in Chapter 5 that the transducer needs to have reasonable radiation efficiency in the frequency range of control. Since the fans generate noise at frequencies as low as 200Hz, the speaker used should have good power handling characteristics, particularly if its diameter is small. Indeed, the radiation efficiency of a baffled speaker is very poor at frequencies whose wavelength is much larger than the speaker driver dimensions. Also, the smaller the cone, the larger its excursion in order to maintain a particular source strength. Indeed, the source strength for a speaker corresponds to its volume velocity which is the piston area times its velocity. The new driver presents a larger power handling capacity than the old one, which is partly due to its larger cone excursion allowed (bigger suspensions, taller voice coil or pole piece).

Finally, the cut-on frequency of the duct for the first cross-mode can be deduced from its geometry based on equation (4.28). The inlet and outlet ducts have a cut-on frequency at 1100 and 1400Hz respectively.

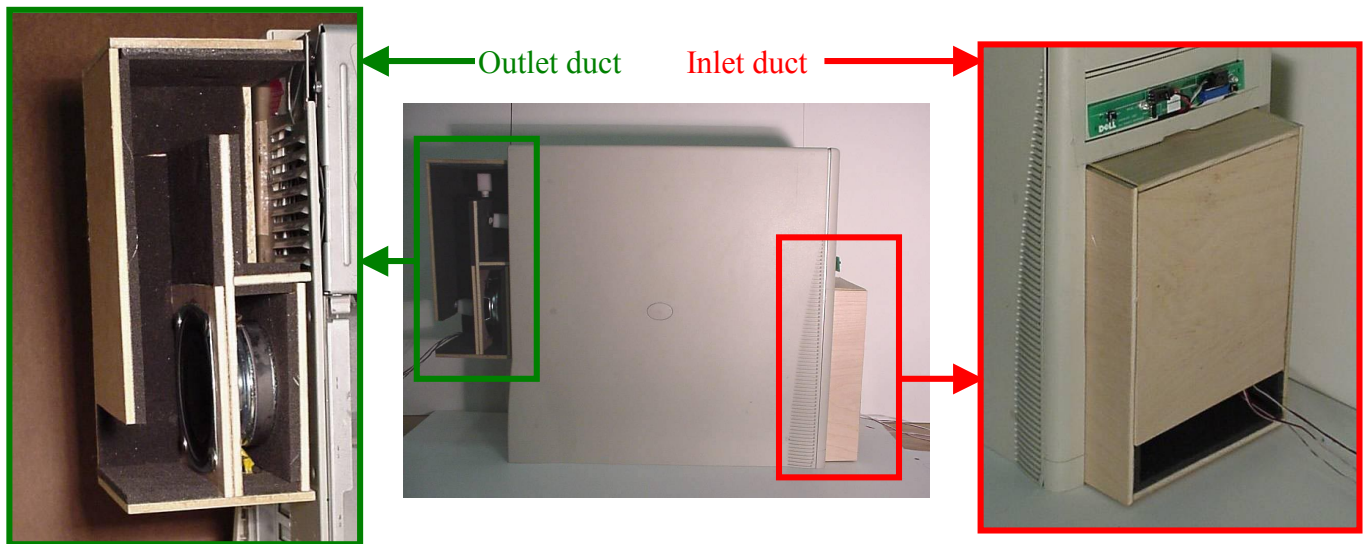


Figure 4.17 Overall view of the PC with the second duct prototype.

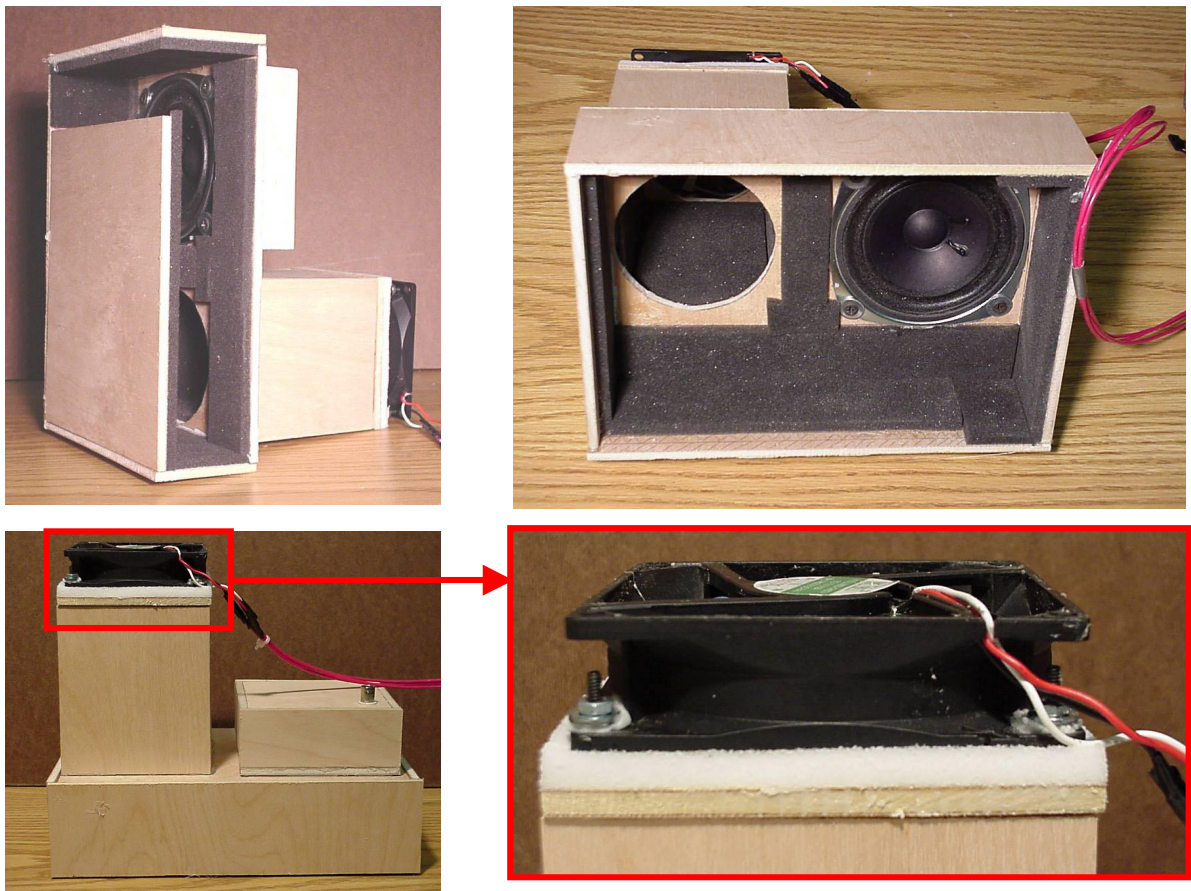
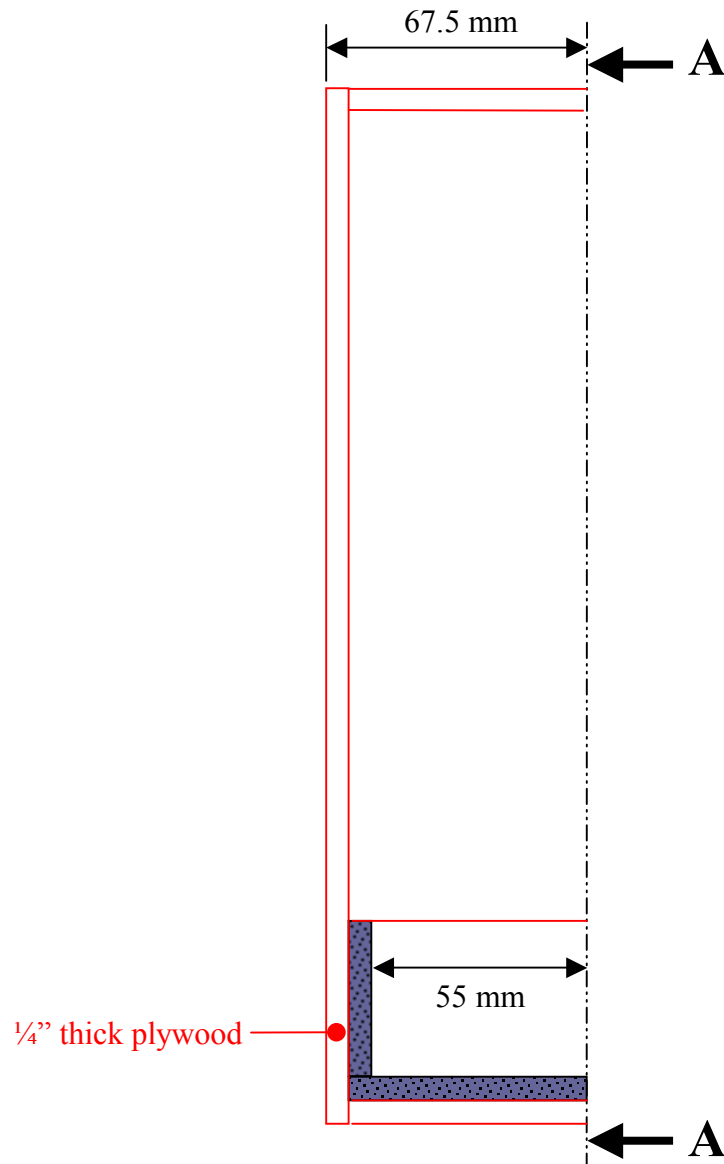


Figure 4.18 View of the inlet duct with the structurally decoupled processor fan.



Rear duct front view

Figure 4.19 Half front view of the second duct prototype (outlet).

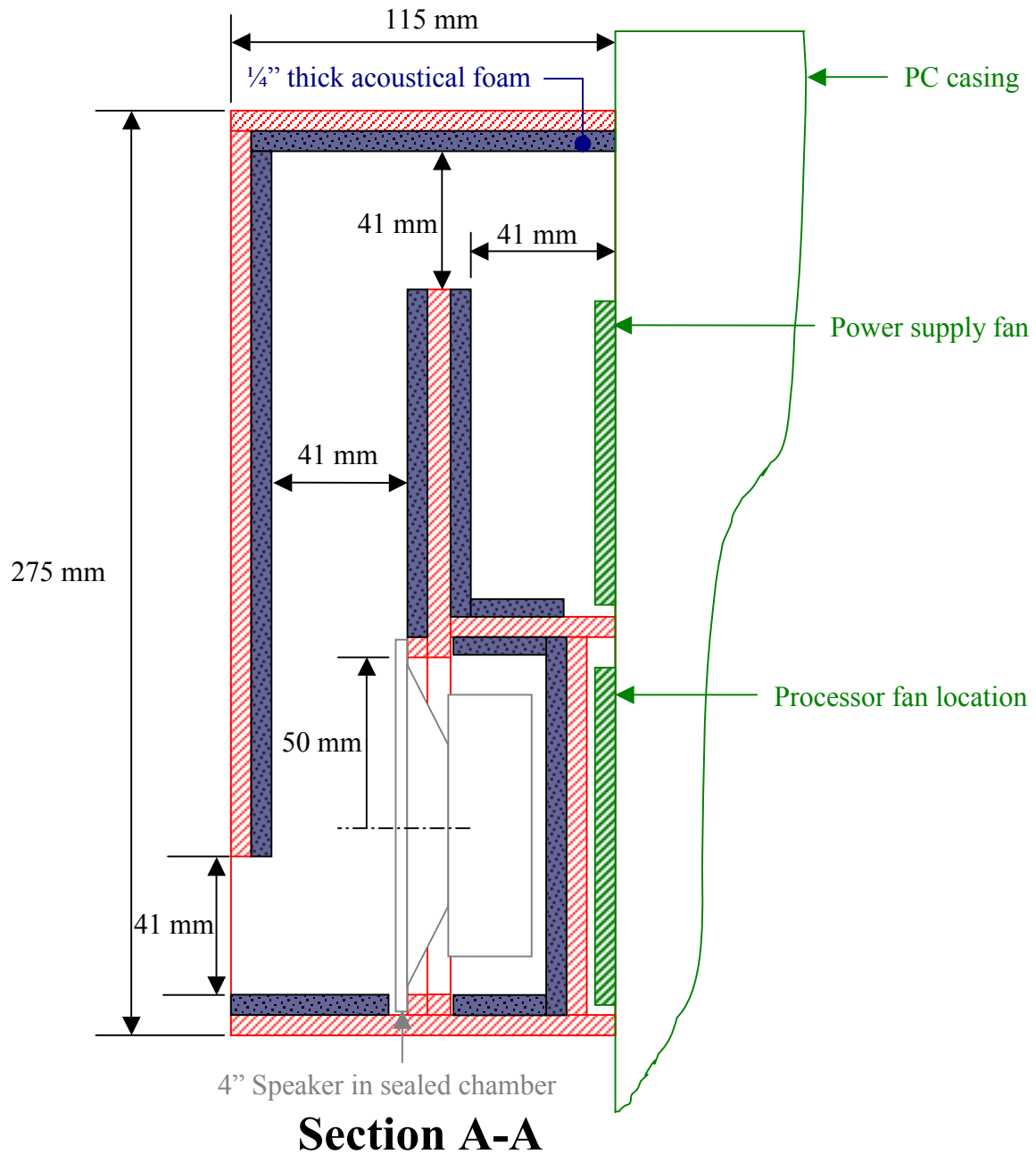


Figure 4.20 Section cut of the second duct prototype (outlet).

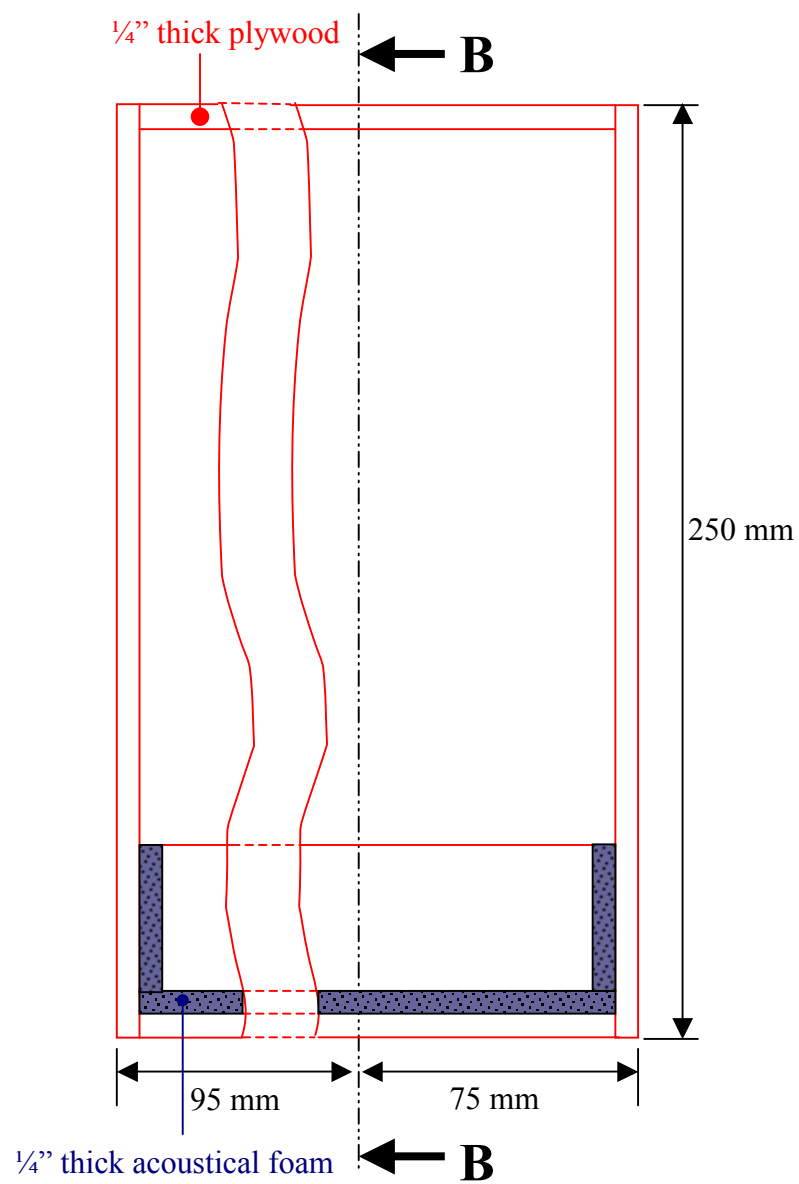


Figure 4.21 Front view of the second duct prototype (inlet).

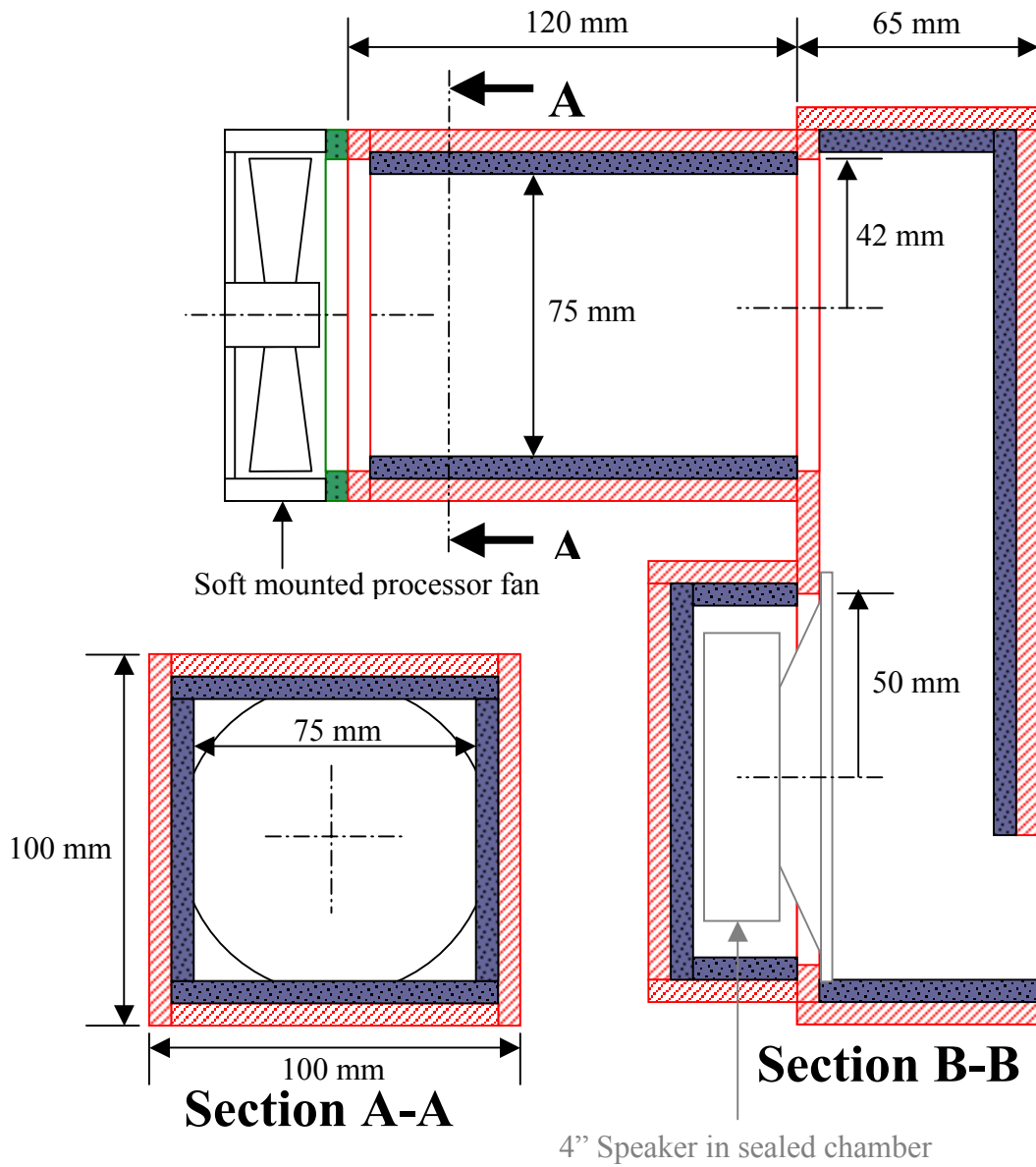


Figure 4.22 Section cut of the second duct prototype (inlet).

4.2.3.2. Acoustical performance

a. Insertion Loss

We have seen in the previous section that acoustic performance was not the leading parameter for the design of the new ducts. Thus, it is very important to now be able to experimentally quantify the new design performance. The insertion loss of the ducts was measured based on sound power level measurements as opposed to the simpler and not quite as accurate sound pressure level measurements performed for the original duct. A 4" diameter speaker in a small sealed enclosure was used as primary sound source. As an approximation to the true insertion loss of the ducts, we measured the sound power level of the sealed speaker when radiating in a half space free field as well as when it was "acoustically loaded" by the inlet or outlet duct. The speaker was fed with a white noise and the sound power level was measured in anechoic chamber using the usual 1 meter radius hemisphere equipped with ten PCB/Acousticel microphones.

The insertion loss measured is shown in Figure 4.23. One first observation is that the new ducts do not perform as well as the first prototype since the attenuation is not better than 25dB up to 5000Hz. Up to 40dB of reduction in sound pressure level were achieved with the first prototype. However, the attenuation starts at 1000Hz, which is as good, if not better, than before. Typically, such a level of performance is still very good and sufficient for our needs.

The second major observation is that, at two low frequencies (200 and 600Hz), the insertion loss is "negative". In other terms, the radiated sound power, at those two particular frequencies, is higher when the speakers are ducted. We believe that this acoustical phenomenon is due to the geometry of the duct. Indeed, let consider a piston "acoustically loaded" by an open-ended duct. Let also assume that the speaker is operating in its piston range, i.e. driven at a frequency lower than the first structural mode of the speaker cone. This is a reasonable assumption for frequencies as low as 200Hz or even 600Hz since such a small cone is expected to present a "cone-breakup frequency" well above 1000Hz. Finally, let observe that the ducts cut-on frequency is above 1000Hz, thus only plane waves are propagating in the frequency range of interest.

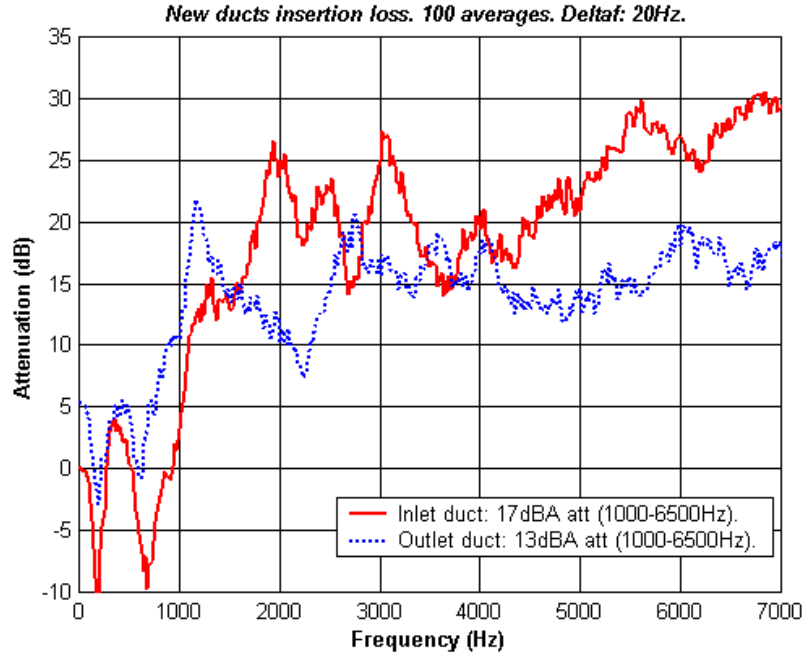


Figure 4.23 New duct prototypes insertion loss.

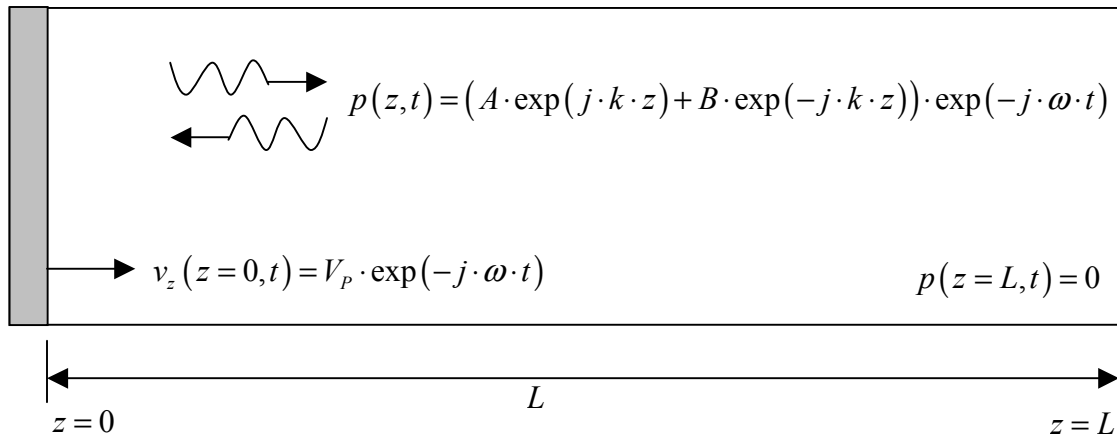


Figure 4.24 Model of the speaker loaded by a duct of length L .

A model of the duct is shown in Figure 4.24. For harmonic excitation, the pressure function is expressed as a sum of waves propagating upstream as well as downstream (there are reflection at both ends of the duct):

$$p(z, t) = (A \cdot \exp(j \cdot k \cdot z) + B \cdot \exp(-j \cdot k \cdot z)) \cdot \exp(-j \cdot \omega \cdot t) \quad (4.38)$$

From Euler's equation, an expression for the velocity can be deduced:

$$v_z(z) = (A \cdot \exp(j \cdot k \cdot z) - B \cdot \exp(-j \cdot k \cdot z)) / \rho_0 \cdot c_0 \quad (4.39)$$

Following the same method as developed at the beginning of this chapter, the boundary conditions in the direction of plane wave propagation can be applied in order to deduce A and B. The ideal boundary conditions shown in Figure 4.24 are a forced particle velocity at $z=0$ (where the speaker is located) and a node of pressure at the outlet (open end at $z=L$). They lead to the following relationship between A and B:

$$v_z(0) = V_p \quad \Rightarrow \quad A - B = \rho_0 \cdot c_0 \cdot V_p \quad (4.40)$$

$$p(z=L) = 0 \quad \Rightarrow \quad A \cdot \exp(j \cdot k \cdot L) + B \cdot \exp(-j \cdot k \cdot L) = 0 \quad (4.41)$$

We can then deduce A and B:

$$A = \frac{\rho_0 \cdot c_0 \cdot V_p}{2 \cdot \cos(k \cdot L)} \cdot \exp(-j \cdot k \cdot L) \quad (4.42)$$

$$B = -\frac{\rho_0 \cdot c_0 \cdot V_p}{2 \cdot \cos(k \cdot L)} \cdot \exp(j \cdot k \cdot L) \quad (4.43)$$

Manipulating the resulting pressure expression we find the solution:

$$p(z, t) = -j \cdot \frac{\rho_0 \cdot c_0 \cdot V_p}{\cos(k \cdot L)} \cdot \sin(k(L - z)) \cdot \exp(-j \cdot \omega \cdot t) \quad (4.44)$$

It is first interesting to observe that the pressure and the velocity are in quadrature, which implies that there is no net flow of energy from the piston surface since the axial intensity is zero everywhere in the duct. We obtain such a result because we did not take into account any damping in the duct and assumed perfectly reflective boundaries. In practice, the open end of the duct has a certain acoustic impedance that will permit transfer of energy to the free field, outside the duct. In such case, there will be a flow of energy in the duct.

However, the most important observation is that the amplitude of the pressure function can go infinite for certain frequencies such that:

$$k_n \cdot L = \frac{2 \cdot \pi \cdot f_n}{c} \cdot L = (2 \cdot n + 1) \cdot \frac{\pi}{2} \quad n = \{0 \quad 1 \quad \dots\} \quad (4.45)$$

We can now have a look at the experimental results for the insertion loss of the ducts. Both ducts present a negative insertion loss at 200 and 600-650Hz. Substituting those values into equation (4.45), a ducted length of approximately 40 cm is deduced, which is close to the actual length of the ducts. Thus, the observed “negative” insertion loss is likely due to the acoustic resonance phenomenon detailed above.

b. Effect on Casing Transmission Loss

After measuring the performance of the ducts, their effect on the PC casing Transmission Loss was quantified. For the following experiments, a small (3.5”) driver, in a sealed enclosure, fed with white noise, was once again used as acoustic source. First, its sound power level was measured in anechoic chamber. Then, it was placed in the PC casing and the new sound power level was measured. The ratio of the two results represents an estimate of the PC casing transmission loss. Finally, the Transmission Loss results for the PC casing, with and without the addition of the lined ducts, were compared. Results are shown in Figure 4.25.

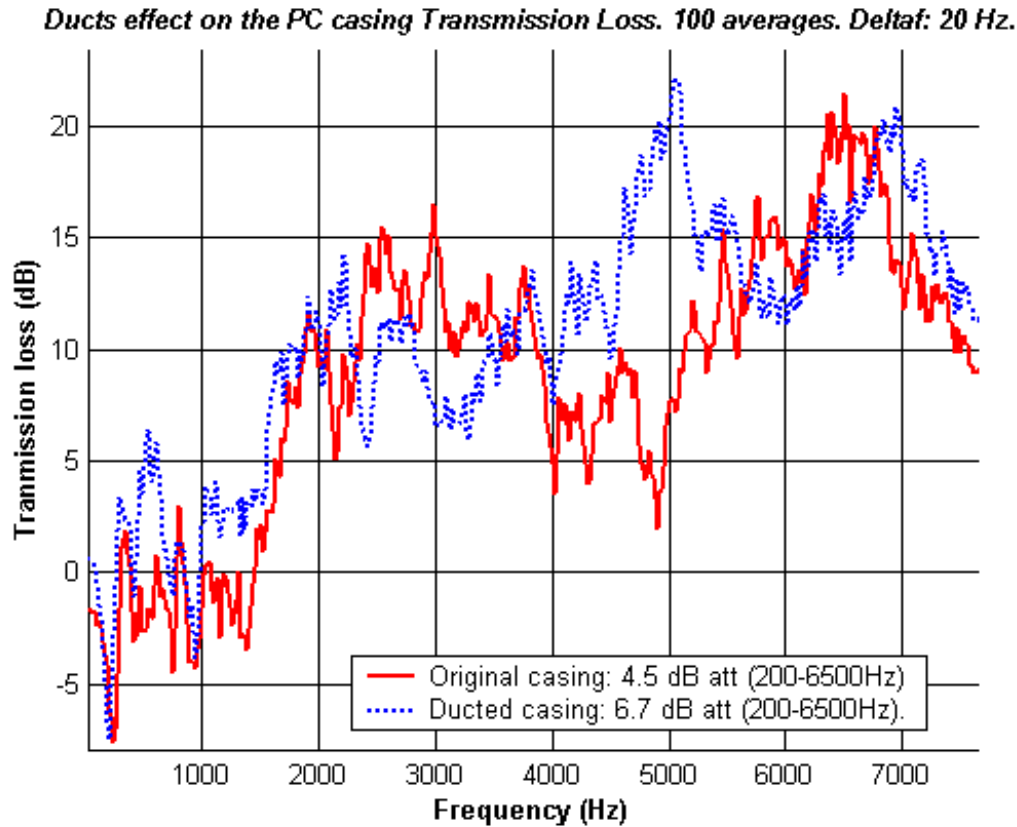


Figure 4.25 Ducts effect on the PC casing transmission loss.

It can first be observed that the Transmission Loss of the original casing is low, never higher than 15dB up to 6000Hz. Below 1500Hz, the casing is acoustically transparent. When, the ducts are added, there is major improvement of the TL between 4 and 5000Hz. Several reasons can explain the limited effect of the PC casing ducting. The first suggestion is that the passive lined ducts performance is not as good as predicted by the Insertion Loss estimate measurements. Considering the liner thickness and large duct dimension (70% open), we could indeed expect the passive absorber to be efficient in a narrow frequency band, rather high frequency. The other major reason could be that the air inlet and outlet are not the only airborne noise radiation paths. A significant portion of the energy inside the PC casing might also be transmitted through the casing panels and various leaks in the assembly. Those paths are under investigation at the time of writing of this thesis.

4.2.3.3. Ducting effect on the PC noise sources.

Previously, the performance of each duct was characterized using a noise source fed with white noise. The goal is now to investigate the effect of the ducts on the noise radiated by the PC. For the following measurements, both the hard drive and processor fan were structurally decoupled from the PC casing in order to focus exclusively on the airborne noise radiation paths that the ducts aim at controlling. The sound power levels measured are always those of the PC, with one or several noise sources operating. The comparisons are made between the “original” casing (no ducts present but processor fan and hard drive structurally decoupled) and the ducted casing (inlet and outlet duct sources structurally decoupled).

The effect of the ducts on the power supply fan noise is shown in Figure 4.26. There is a significant attenuation of both the broadband and harmonic noise generated by the fan over the whole frequency range of radiation. Since we achieve a significant attenuation at low frequencies (below 1000Hz), we believe that the duct is an effective reactive silencer because the liner is not supposed to be efficient at such low frequencies. We also observe that the fan blade passing frequency slightly decreases (from 250 to about 200Hz) when the ducts are added. This might be due to an increase in airflow resistance due to the addition of both ducts, thus decreasing the fan speed.

The processor fan is an other noise source whose sound power level might be reduced by the presence of the ducts. As part of the ducting, the processor fan has been relocated inside the casing and is attached to the inlet duct. Also, it is structurally decoupled from the duct. In Figure 4.27, the sound power level radiated by the ducted PC is compared to the non-ducted case (still with the processor fan decoupled). We obtain the same results as observed with the power supply fan. The multiple high frequency tones seen on the red curve are actually electrical noise (60Hz) as observed at the beginning of this chapter.

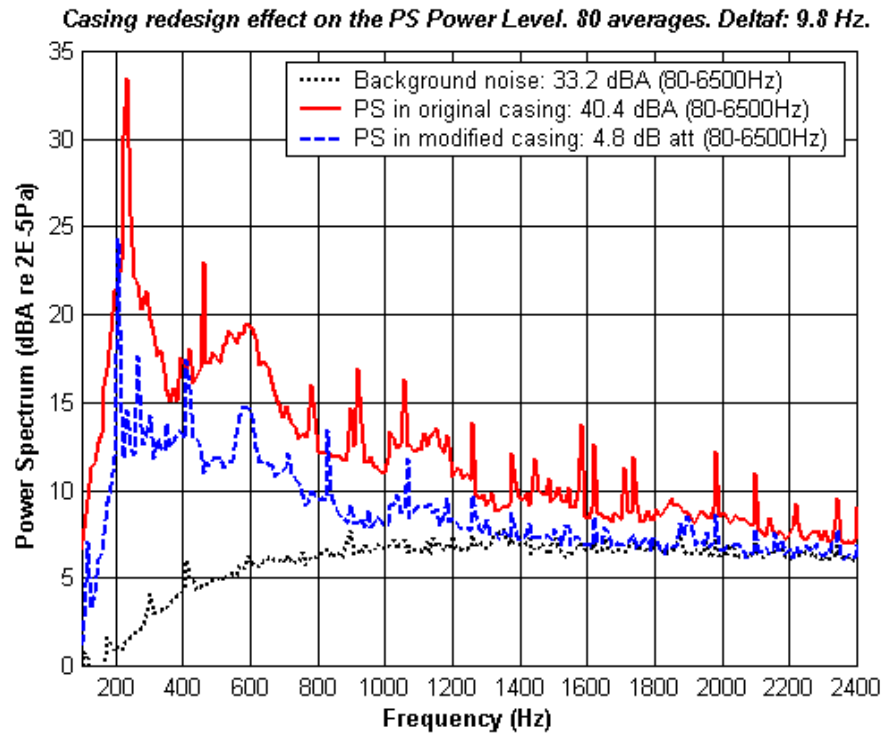


Figure 4.26 Ducts effect on the power supply fan sound power level.

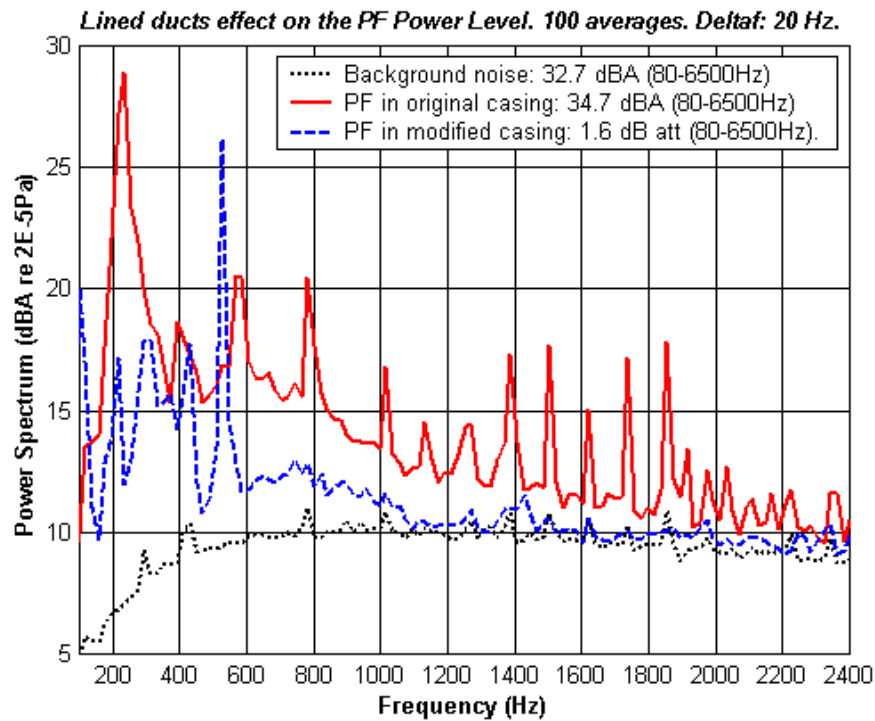


Figure 4.27 Ducts effect on the processor fan sound power level.

Finally, the sound power level radiated by the hard drive when mounted in the original casing (but structurally decoupled) was compared to the sound power level when the ducts are added. The results are shown in Figure 4.28. In that case, the ducts do not induce an important drop in the sound power level. The main difference with the fans is that the hard drive is radiating in the PC casing for both ducted and non-ducted configurations. The fans radiate originally in a free field and then in the duct. Thus the ducts do not modify the hard drive noise radiation properties, contrary to the fans. The results below show that we either the ducts have a limited true insertion loss or that the two other airborne noise radiation paths, which are the sound transmission through the casing panels and the leaks in the assembly, both not currently addressed, are significant.

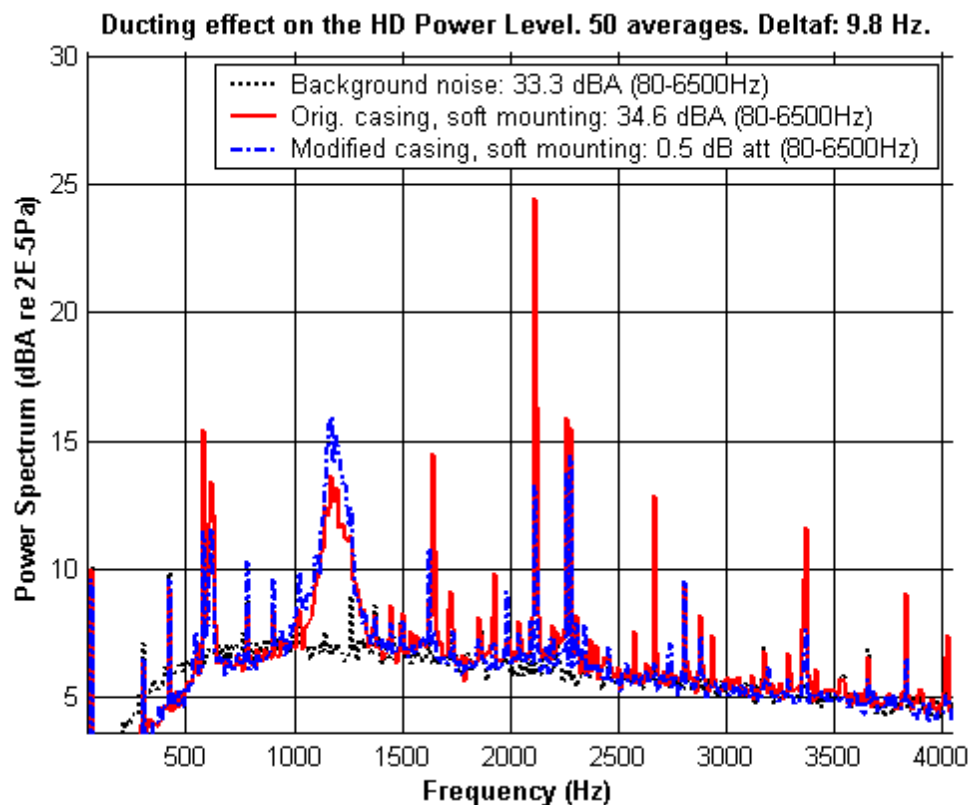


Figure 4.28 Ducts effect on the hard drive sound power level.

4.2.3.4. Overall passive treatment performance.

We can finally show the combined effect of all the passive noise control treatments applied on the PC. Two of the noise sources were structurally decoupled from the PC casing to reduce the structure-borne noise radiation path. These devices were the chassis mounted processor fan and the hard drive. We also tried to suppress part of the airborne noise radiation paths by ducting both the PC air inlet and outlet. The change in sound power level radiated by the three stationary noises sources mounted in the PC is shown in Figure 4.29.

The combined effects of the different treatments result in a reduction of most of the tonal content of the PC noise. The resulting noise is mainly low frequency (below 1000Hz) with some hard drive noise at higher frequencies (between 1000 and 2500Hz). The power supply fan still radiates an important broadband noise accompanied with the blade passing frequency tone at 250Hz. This can be due to either a lack of performance of the lined ducts at lower frequencies (below 1000Hz) or to the presence of other airborne noise radiation paths such as transmission through the casing panels and assembly leaks. But the major hypothesis is that the power supply block fan radiates energy through a structure borne path, which has not been investigated at the time of writing of this thesis.

Due to time limitations, the work performed on passive noise control treatments was halted here. Our next goal is to investigate the feasibility of Active Noise Control in ducts, which constitutes the theme of Chapter 5.

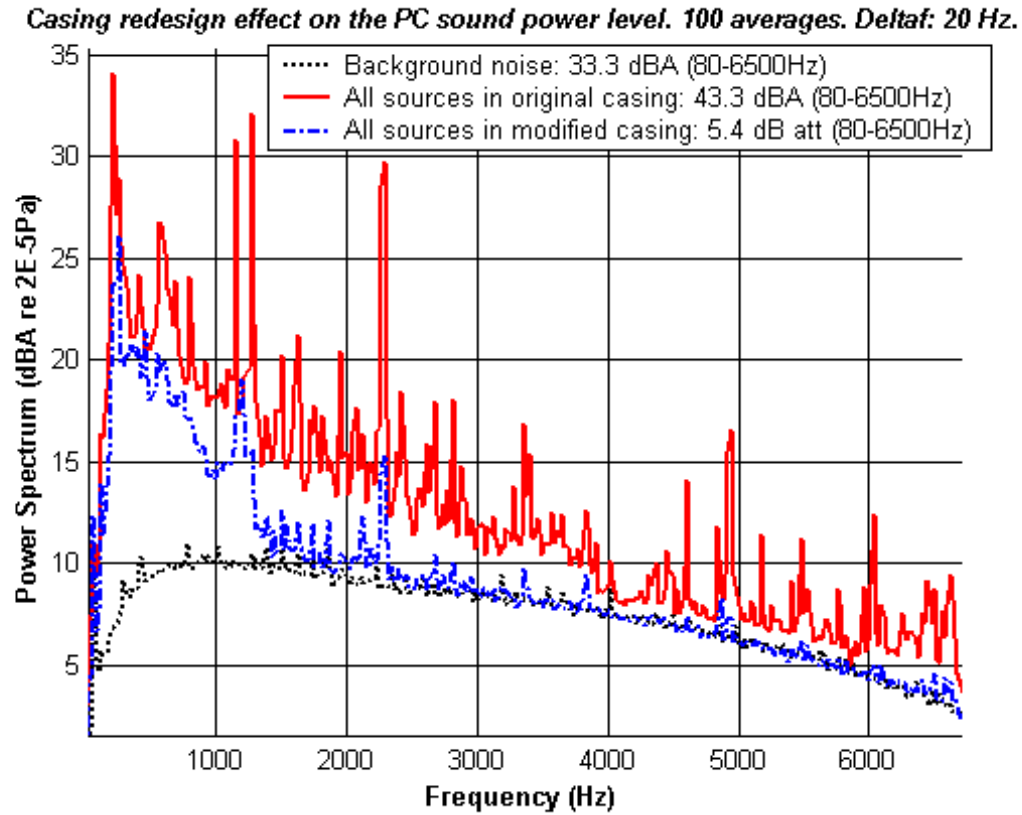


Figure 4.29 Overall passive treatment effect on the sound power level of the PC.

Chapter 5. Global Active Noise Control

As its name suggests, the global noise control methodology aims at controlling a primary noise source radiation at all points in its surrounding space. Usually, this requires the “creation” of a secondary noise source of similar complexity as the primary source [9]. Such canceling noise source will then typically be located in the vicinity of the primary source and fed with the same signal, but out of phase. This will typically results in noise cancellation.

A typical example is the case of two loudspeakers located next to each other. If the relative polarity of the speakers is inverted and they are fed with the same signal, one will then hear a significant loss in bass. Indeed, at frequencies where the acoustic wavelength is much larger than the loudspeakers spacing, there is effective global “noise cancellation”.

Compared to the local control strategy that targets only noise reduction at specific points in space, the global control strategy is a preferred method when the cavity in which the sources reside is to be quiet, regardless of the user position. The only case where local control is preferred is when the user will not change position. Although in most situations one might prefer the global control solution, it will be shown in this chapter that it sometimes requires significant redesign of the system in order to be efficiently implemented. Moreover, the global control strategy presents important implementation constraints such as causality.

Active Noise Control systems vary widely in terms of complexity, from single input / single output to multiple channel controllers. The goal here is to design the simplest Active Noise Control system that will be not only cost effective, but also require limited space to be implemented in a typical PC. In this chapter, we will first justify the use of ducts to implement the global ANC system. We will then represent the typical architecture for a feedforward ANC system and describe some of the key parameters that ensure a good performance. In a second part, the systems designed for the current project will be described and their performance quantified.

5.1. Review of the Filtered-X LMS Feedforward Control System

In section 2.3.1, adaptive digital filters were introduced along with one implementation case in the form of a SISO feedforward controller. Different algorithms are available to perform the actual update of the adaptive filters. Most active noise controllers used in VAL are based on the Least Mean Square algorithm. For our practical designs, we will use the Filtered-X LMS algorithm, that is more robust than the LMS algorithm in real life systems with secondary paths [9] as detailed in section 2.3.2.1.c.

The block diagram of a Filtered-X digital feedforward controller is shown in Figure 5.1. The control objective is to minimize the error signal $e[n]$ that is a sum of the disturbance signal $d[n]$ generated by the unknown plant $P[z]$ and the signal $y'[n]$ generated by the controller.

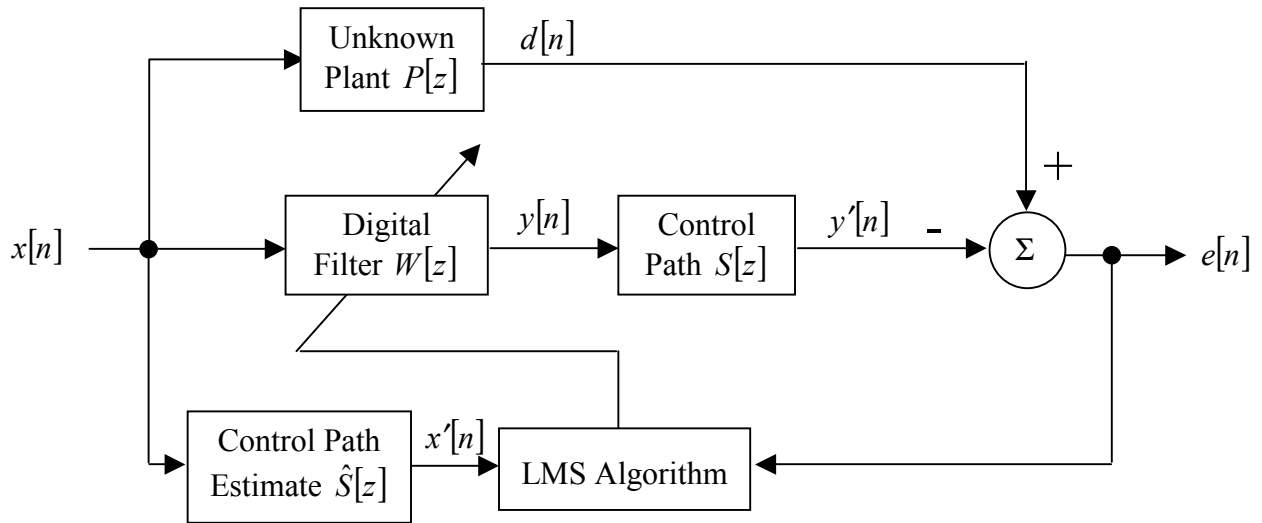


Figure 5.1 Single channel feedforward controller based on the Filtered-X LMS algorithm

The digital filter $W[z]$ is consisting of L taps and used to generate a control signal in real time. It filters a reference signal $x[n]$ that is assumed to be well correlated with the disturbance $d[n]$. The filter weights are continuously updated in order to minimize the cost function $\xi[n] = E[e^2[n]]$. The Least Mean Square algorithm used for this purpose is a “simplified algorithm” since it does not minimize the actual cost function but an estimate which is the instantaneous squared error $e^2[n]$. This gradient-based algorithm has been introduced in the section 2.3.1.3.b. Thus, we will here, directly express the relation for the control filter weights update:

$$w_l[n+1] = w_l[n] + \mu \cdot x'[n-l] \cdot e[n] \quad l = \{0, 1, \dots, L-1\} \quad (5.1)$$

where:

$$x'[n-l] = \sum_{k=0}^{K-1} \hat{s}_k \cdot x[n-l-k], \quad (5.2)$$

with $\hat{s}_k$ being the coefficients of the digital filter of K taps modeling the secondary path:

$$\hat{S}[z] = \sum_{k=0}^{K-1} \hat{s}_k \cdot z^{-k} \approx S[z] \quad (5.3)$$

In equation (5.1), μ is called the convergence coefficient. It determines the rate at which the algorithm is seeking the optimal solution. As we have seen in section 2.3.1.3.b, the maximum value for the convergence coefficient μ , in the case of Filtered-X LMS algorithm, is a function of the filtered reference signal power $E[x'^2[n]]$, adaptive filter size L , and overall delay Δ in the secondary path, in samples. The major limiting factor is usually the delay in the secondary path:

$$\mu_{\max} = \frac{1}{L \cdot \Delta \cdot E[x'^2[n]]} \quad (5.4)$$

5.2. Implementation of ANC in Ducts

5.2.1. Ducts: a Privileged Global ANC Environment

It was shown in section 4.2.1.1.c that, below a certain critical frequency that is a function of a duct cross-sectional dimension, only plane waves are propagating in a waveguide. This is the simplest form of sound field for which the pressure fluctuations are only a function of time and distance in the propagating direction (constant amplitude in the plane of propagation).

We consider a speaker mounted on one wall of a rectangular duct and focus on frequencies below the duct cut-on frequency. In that case, the loudspeaker will be able to generate plane waves propagating upstream and downstream the duct. Higher order modes should only be excited in the near field of the loudspeaker and decaying exponentially as we move away from the speaker. Neglecting those near field effects, the speaker can be modeled as a plane source. By driving the loudspeaker with the appropriate source strength q_s and phase, downstream propagating disturbances due to the primary source of strength q_p can be cancelled as shown in Figure 5.2.

For such a cancellation strategy, it can be easily shown that the primary and secondary sources strengths are simply related:

$$q_s = -q_p \cdot \exp(-j \cdot k \cdot L) \quad (5.5)$$

The time domain relationship is given by:

$$q_s(t) = -q_p \left(t - \frac{L}{c_0} \right) \quad (5.6)$$

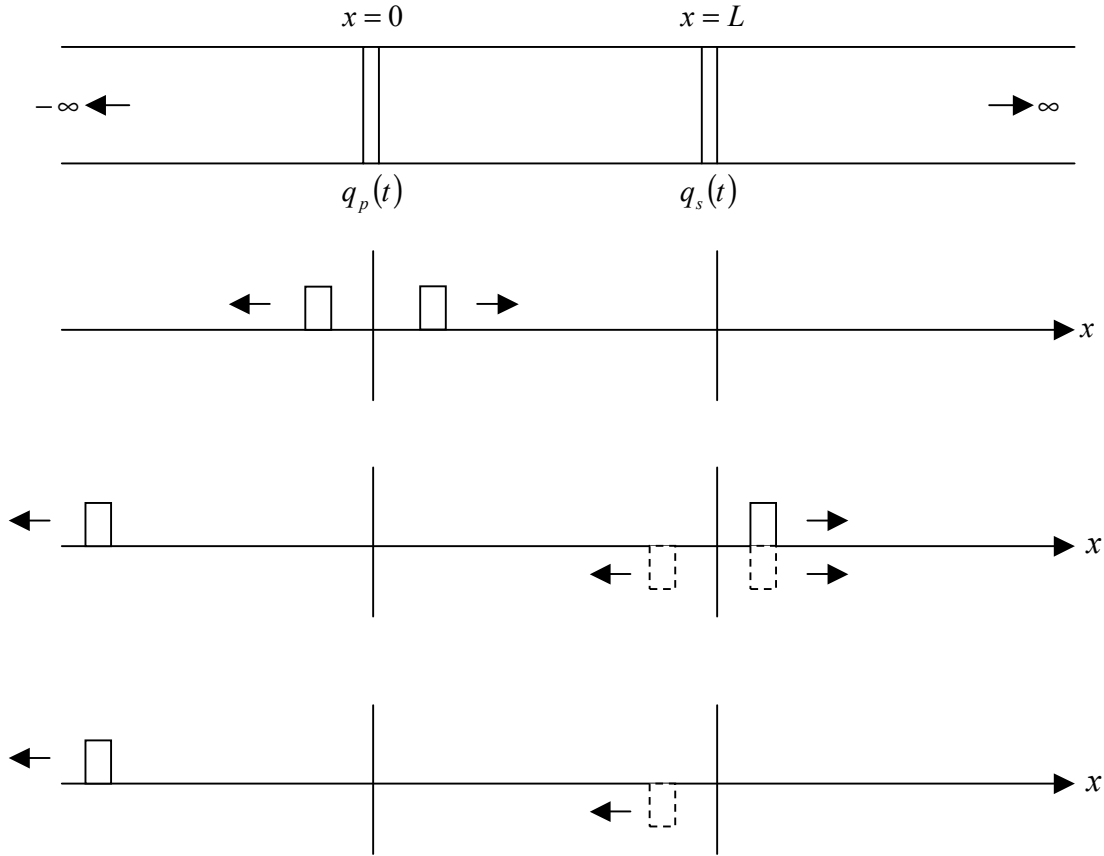


Figure 5.2 Secondary source controlling a propagating disturbance in duct.

In order to cancel radiation downstream of the secondary source, its source strength must be similar to the primary source, but of opposite sign and delayed by a time L/c_0 , which corresponds to the acoustic wave propagation delay from the primary source to the secondary source.

Thus, a single microphone, located upstream, could be used to obtain a reference signal for a single input / single output controller that would drive a single secondary noise source in order to minimize the pressure level in a downstream section. Since the disturbance is a simple plane wave, only one sensor in the cross-section is necessary to accurately sense the medium. Also, the secondary noise source will not need to generate a complex noise field. Thus, a single speaker is enough. We will now review in more detail the architecture of the SISO feedforward LMS digital controller for use in ducts.

5.2.2. ANC Architecture

For the Active Noise Control of plane waves propagating in a duct, we will build a feedforward digital controller as shown in Figure 5.3. The primary noise source, located upstream, generates a disturbance propagating downstream that is sensed by a reference microphone. We try to minimize the level at the error microphone that is located downstream. The reference signal is used to generate a control signal that will be fed to the speaker located between the reference and error sensors. Such a controller is referred as SISO (single input / single output) since a unique reference signal is used to drive a single secondary noise source to minimize the level of a unique error sensor.

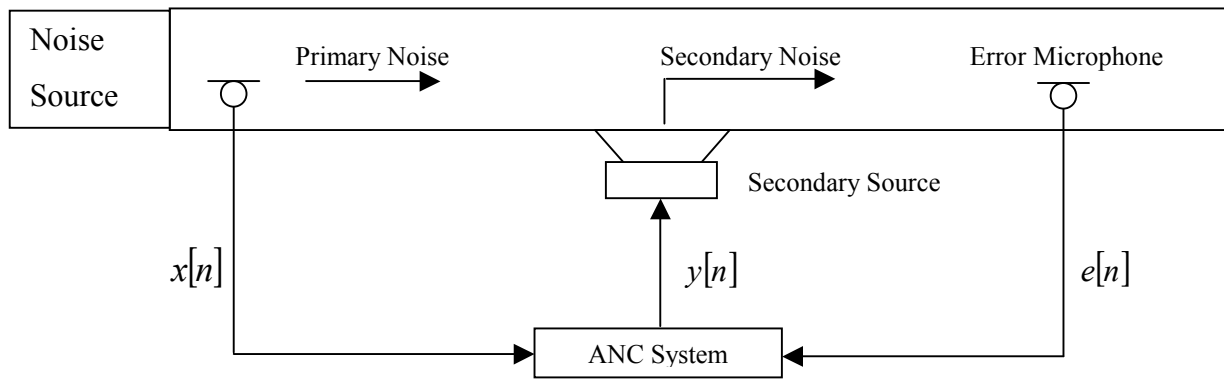


Figure 5.3 Single channel broadband feedforward ANC in duct

For such a controller, we can represent a more complete block diagram representing the overall system. We can distinguish three domains with different types of signals: acoustical, electrical (analog) and digital domains. Transducers that convert pressure fluctuations into electrical signals (and vice versa) are used. These electrical signals are then converted to discrete sequences in order to be able to be analyzed and processed by the digital controller. Similarly, the controller output sequence has to be converted back to an analog electrical signal that can be fed to the appropriate actuator. The block diagram of the complete controller is shown in Figure 5.4.

We define the direct acoustical path from the reference sensor to the control speaker as R2C (reference to control) and from the control speaker to the error sensor as C2E (control to error). Apart from the signal conditioners and amplifiers used to drive the different transducers, we also observe the presence of analog low-pass filters that are used to prevent aliasing effect and filter out quantization noise.

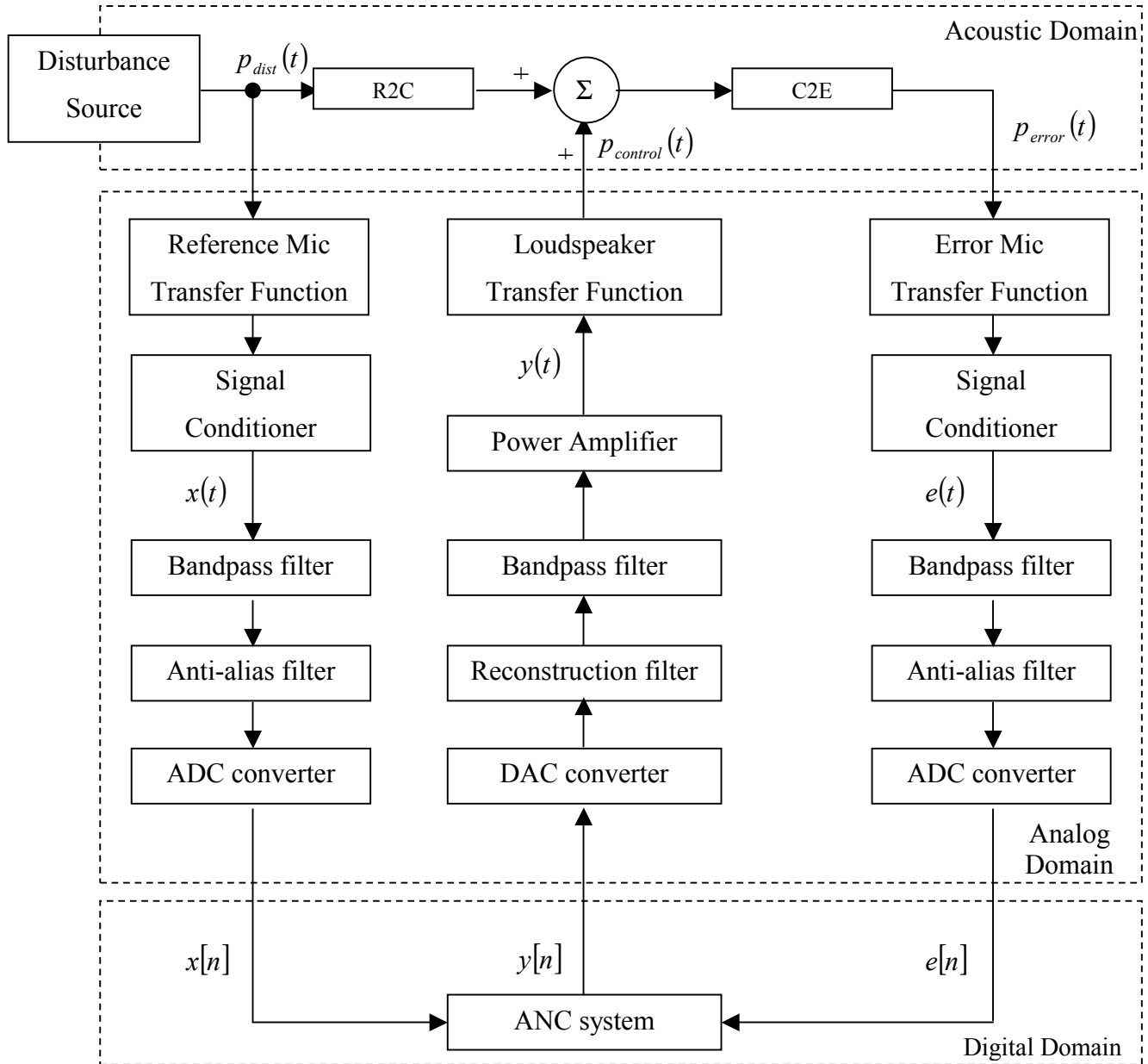


Figure 5.4 Complete block diagram of a SISO feedforward digital controller

Since we always limit the frequency range of control, bandpass filters can also be added to the various paths. They can eventually, replace the reconstruction and anti-alias filters that are only lowpass type. However, it will be shown below that, due to causality issue, the use of filters in the control system introduces delays that can interfere with the broadband control performance. Thus, their use is minimized.

5.2.3. Key Parameters for the Controller Performance

5.2.3.1. Reference Signal Selection - Coherence

One of the key parameters for feedforward Active Noise Control is the selection of an appropriate reference signal. The reference signal is used to obtain information regarding the disturbance signal before it reaches the secondary source. In order to properly control the disturbance signal, the most complete information about its characteristics needs to be determined using the appropriate sensor.

Once the appropriate reference sensor is selected, a second important task is to ensure that the detection and error sensors signals are not largely corrupted by measurement noise. In section 2.3.2.1.d, we have shown that measurement noise at the detection sensor affects the performance of the controller since it will then perform a balance between canceling the acoustic noise and reducing the amplification of measurement noise at the detection sensor. We can show again the electrical block diagram of a single channel control system to illustrate the presence of measurement noise in real life systems (see Figure 5.5).

It was shown in the section 2.3 that, when there is no measurement noise, the frequency response function of the optimal controller is:

$$W^o[z] = \frac{P[z]}{S[z]}, \quad (5.7)$$

while, in presence of noise, it is given by:

$$W^o[z] = \frac{P[z]}{S[z]} \cdot \frac{1}{1 + 1/\rho[z]}, \quad (5.8)$$

where $\rho[z]$ the signal to noise ratio at the detection sensor. Thus, the optimal filter is reduced in presence of noise at the detection sensor.

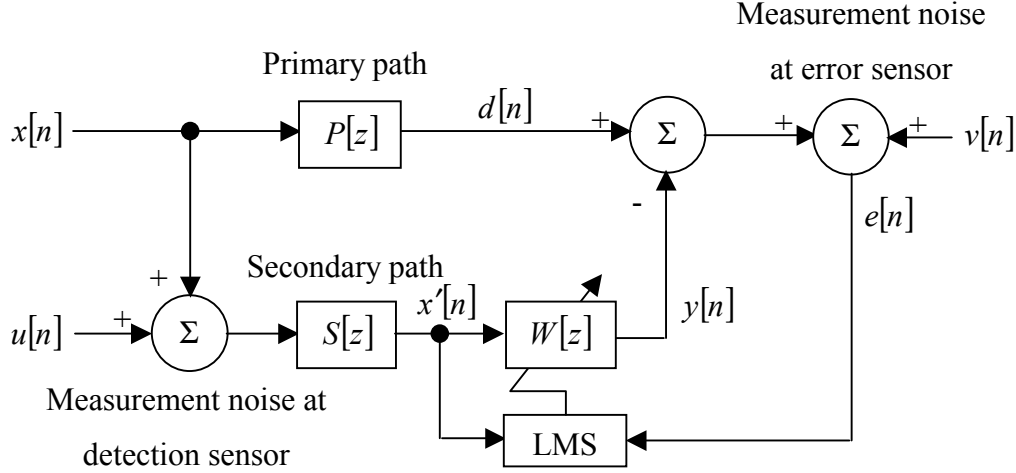


Figure 5.5 Block diagram of ANC system with noise on the input signals

If we define $S_{dd}(\omega)$ and $S_{ee}(\omega)_{\min}$ as the power spectral densities of the error signal before and after control respectively, it can be proved that they are related by the ordinary coherence function $\gamma_{ud}^2(\omega)$ between the outputs of the detection and error sensors:

$$\frac{S_{ee}(\omega)_{\min}}{S_{dd}(\omega)} = 1 - \gamma_{ud}^2(\omega) \quad (5.9)$$

This result that was derived by Ross (1980) constitutes a convenient tool for estimating the performance of an active noise control system based on simple measurements. Such an expression shows that the reference and error signals should be highly correlated to ensure proper control. If the reference sensor signal is poorly correlated with the disturbance signal, then the coherence function between reference and error signals will ultimately be low which translates into a reduced controller performance.

For those reasons, one of the important tasks in the controller implementation phase is to select an appropriate reference signal and ensure that none of the measured signals are corrupted by measurement noise. In our application, since we want to control the broadband noise radiation from the ducted fan, we will show that a microphone is the adequate sensor. We could use a simple tachometer signal or accelerometer on the fan for controlling pure tones (BPF and its harmonics). But such signals will not be correlated with airflow-induced noise for example. The advantage of an accelerometer or a tachometer is that, contrary to a reference microphone, they are not sensitive to the control signal and thus cannot generate destabilizing feedback loops in the control system [24].

5.2.3.2. Causality

The second key parameter in the controller design is to ensure that the control be causal. For the problem of Active Noise Control in a duct, the causality issue is quite easy to illustrate. We assume a disturbance propagating from the primary source to the secondary source location at the phase speed c . The time it will take for this disturbance to reach the secondary source location is proportional to the distance d between primary and secondary noise sources. $\delta_p = \frac{d}{c}$ is called the transport delay in the R2C (reference to control) acoustical path.

A certain delay also exists in the control path, which starts at the reference sensor and ends at the control speaker, passing through the controller. This delay is due to the transducers, the analog and digital filters present in the path, the ADC and DAC converter as well as the controller itself. We will call this control path delay δ_c . In order to build a causal control filter, the total delay in the control path should be less than that in the direct acoustic path.

The causality problem is an issue only in the control of random noise that is by nature completely unpredictable. In case of control of pure tones such as the fan BPF and harmonics, those signals being perfectly predictable, the need for a delay in the direct path is eliminated.

In order to make the noise control system causal, the distance between reference sensor and control transducer will have to be maximized. Also, the different delays in the controller path should be minimized. It will be shown that, by over sampling the controller system, we could minimize the delays in the digital controller, ADC and DAC converters as well as in the analog filters.

5.3. Practical Design

During this project, two prototypes of lined ducts with integrated Active Noise Control system were built. In section 4.2, both designs were presented, but the analysis was limited to the passive properties of the ducts. In this section, the geometry of each duct is briefly reviewed, locating the transducers used in ANC. The control setup as well as the results achieved for both prototypes are then presented in detail.

5.3.1. First Duct Prototype

5.3.1.1. Control System Presentation

a. Duct Geometry and Transducers

The geometry of the first duct prototype is shown in Figure 5.6, illustrating the typical location of the reference and error microphones used for the control. Nevertheless, the reference signal used was not always a microphone. Moreover, the reference microphone would be located at different upstream locations closer to the fans air outlet. On the other hand, the secondary source and error sensor always remained at the same location.

Depending upon the reference microphone location, the acoustical path between reference microphone and secondary source varies between 160 and 200mm, which corresponds to lags of 0.5 to 0.6 ms. This lag is an important parameter for the controller since it constitutes the maximal delay that can be tolerated in the control path.

As was described earlier, the secondary noise source is a 3" diameter speaker in sealed enclosure. The choice was made based on size and weight limitations rather than actuator performance. The driver used has little authority in the frequency range of ANC due to its size. Unfortunately, the lack of efficiency at low frequencies is not compensated by a large power output since the diaphragm excursion is very limited. It will be shown that this had a negative impact on the control results.

Several types of reference sensors were tested. Firstly, we used accelerometers mounted on or near the vibrating sources. An accelerometer could also be attached on the processor fan directly. For the power supply block fan, the accelerometer was mounted on the power supply block since the fan itself was not directly accessible. The fan being rigidly mounted on the power supply block, we could still obtain a reference signal that was highly correlated to the fan vibration.

The pressure sensors originally used were B&K high precision microphones. We rapidly switched to more compact and affordable Acoustical/ PCB microphones since these yielded similar control results. Indeed, even though the PCB microphones have a much lower signal to noise ratio than the B&K's, the sound pressure levels measured in the duct were large enough to allow us to use low cost PCB sensors.

b. Controller Settings

The frequency range of Active Noise Control is limited for several reasons. One of them was developed earlier in this chapter: it concerns the noise source complexity. The control is limited to plane waves disturbances. For the first duct prototype, the cut-on frequency is 1600Hz. We will thus not actively control the noise radiated above that frequency.

In practice, the wider the frequency range of control, the higher the controller sample rate since it has to be at least twice the highest frequency controlled so that the conversion from analog to digital domains is done properly (i.e., without aliasing phenomenon). Consider a digital filter modeling an impulse response of duration T . Since $F_{\text{samp}} = N/T$, the higher the controller sample rate F_{samp} , the more taps N in the digital filter are needed to model the impulse response in the discrete domain. Thus, the control bandwidth can also be reduced due to limited computational resources.

We also limit the low frequency extension of active noise control. One reason is that the primary noise source does not radiate any noise below a certain frequency (typically 100Hz for the PC fans). Also, as was said above, the secondary noise source has very little authority at low frequencies. The controller will compensate for the weakness of the loudspeaker to a certain extent. Eventually, it will saturate if it cannot provide enough gain for the compensation. In order not to saturate the noise controller, the low frequency content of the input signals to the controller will be filtered out.

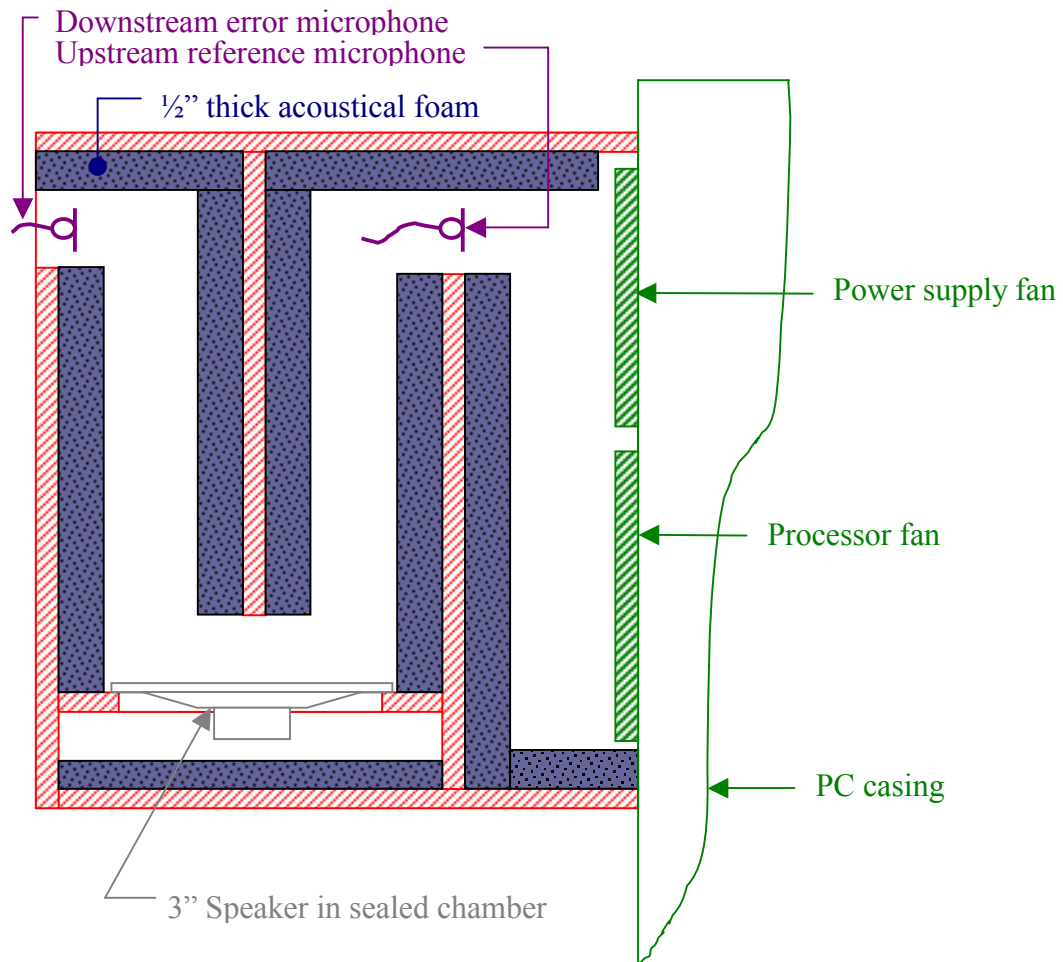


Figure 5.6 First duct prototype.

5.3.1.2. Reference Signal Investigation

The first work carried on the active noise control system was to select the appropriate reference signal that would provide complete upstream information about the propagating disturbance. In the following experiments, both the power supply block fan and processor fan are operating. As we have seen before, the performance of the active noise controller can be predicted by measuring the simple coherence function between reference and error sensor (c.f. equation (5.9)). Indeed, the error signal content that is correlated with the reference signal will be suppressed when the controller is operating normally.

a. Vibration Versus Pressure Reference Sensor

In Figure 5.7, we plot the power spectrum of the error microphone signal. We simulate the effects of a single channel feedforward ANC system minimizing the pressure level at the error microphone location based on the information provided by a reference signal. One of the reference sensors is an accelerometer mounted on the PC casing where as the other is a microphone located upstream near the power supply and processor fans. Figure 5.8 shows the achievable attenuation as well as the simple coherence function between the reference signals and the error signal.

Both sensors provide a very different information about the primary noise source. The accelerometer signal is exclusively correlated with the pure tones radiated by the fans. In such case, the coherence function is close to zero over the whole frequency range except at discrete frequencies. Some frequencies correspond to the fans BPF and harmonics. We also observed in section 3.2.2.2 that the multiple pure tones radiated by the processor fan, between 1000 and 1500Hz, were due to structural excitation of the PC casing by the fan. This behavior is well shown here since the accelerometer signal is highly correlated to the sound pressure level at those discrete frequencies as well.

On the other hand, the reference pressure sensor signal is well correlated with the error signal over a broad frequency range. This is logical since the broadband noise radiated by the fan is mainly airflow induced. The tonal content of the radiated noise can also be controlled using a pressure sensor as reference. The power supply block fan BPF can even be better attenuated using a microphone rather than an accelerometer as reference sensor. On the other hand, the multiple tones due to structural excitation of the PC casing by the processor fan are better attenuated when using a vibration sensor as reference.

In terms of overall noise control, the pressure sensor constitutes a favored reference since it leads to a theoretical attenuation of at least 8 dBA over the 200-1500Hz frequency range, and more than 20dB at the power supply block fan BPF (250Hz). The accelerometer does not provide any broadband noise control but can lead to an attenuation of 5 to 15dB of the tonal content of the radiated noise which translates into an overall noise reduction of almost 4dBA in the 200-1500Hz frequency range.

Reference signal selection in first duct. Pressure versus vibration sensor. Deltaf= 4Hz

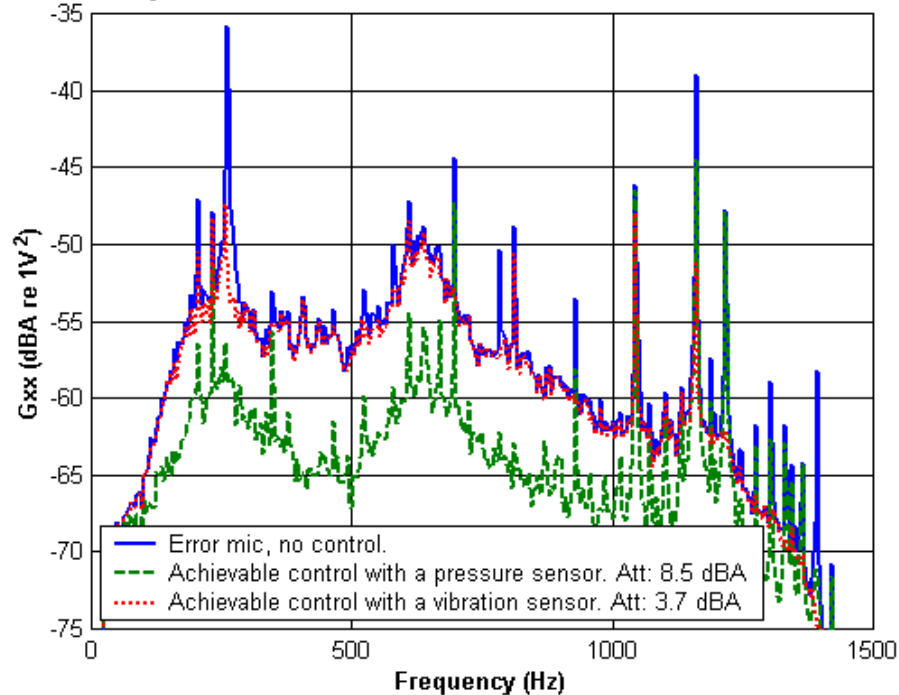


Figure 5.7 Single channel ANC in duct. Reference signal type selection.

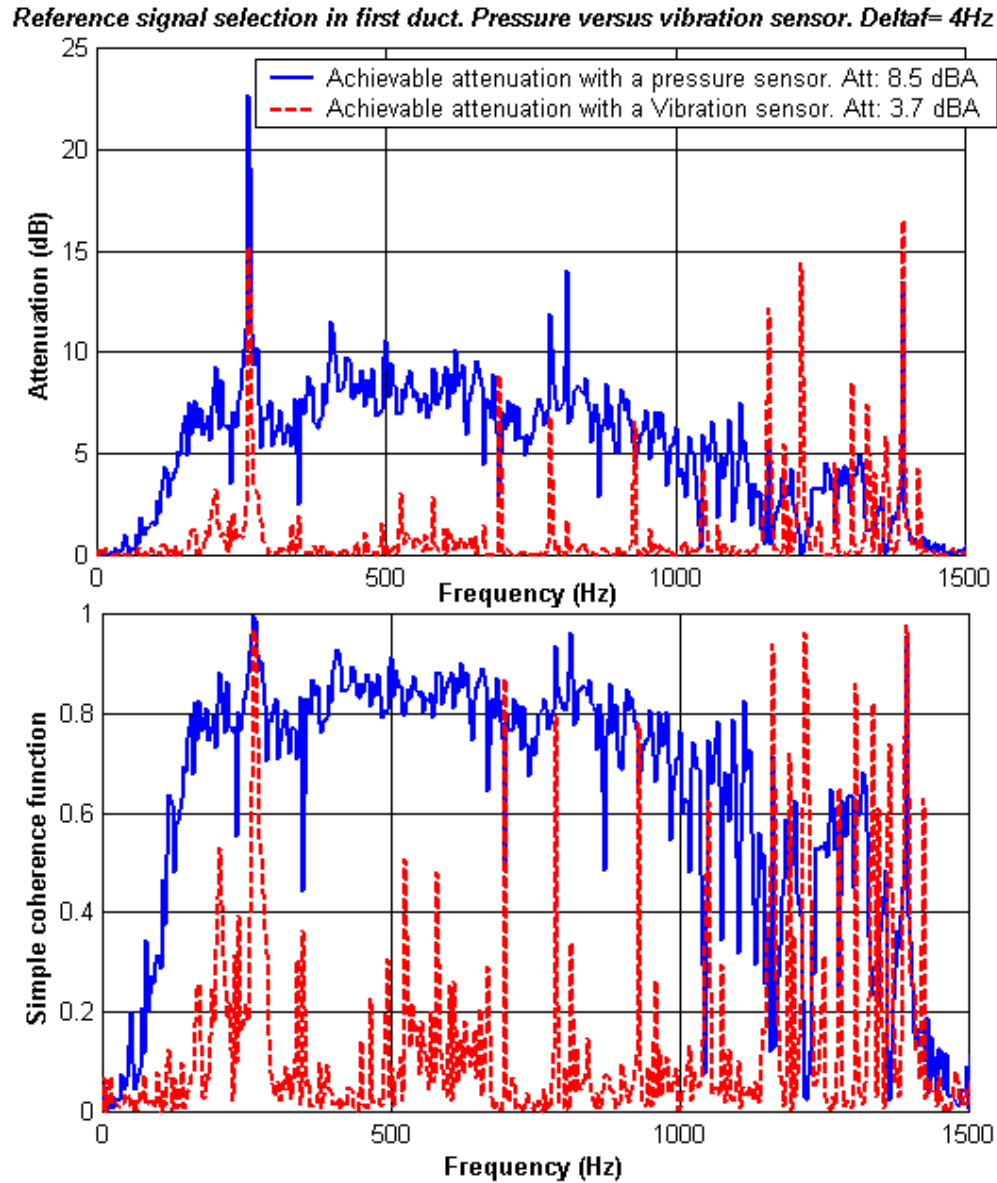


Figure 5.8 Single channel ANC in duct. Reference signal selection (concluded).

For the single channel feedforward noise control strategy in ducts, we will primarily use a pressure sensor as reference since it leads to the largest broadband noise attenuation. Ideally, a pressure and a vibration sensor should be combined as references. This can be done approximately by using a microphone rigidly attached to the PC casing: it can then behave not only as a microphone but also as an accelerometer.

b. Microphone Location Effect

The reference microphone can be placed at various locations in the duct. We here want to test the effect of the microphone location on the achievable noise control. The same experiments as above are performed, but using only a microphone as reference sensor. Four different locations are investigated:

- Location 1: Near the power supply block fan, airflow side (microphone in duct)
- Location 2: Near the processor fan, airflow side (microphone in duct)
- Location 3: Near the power supply block fan, microphone in PC casing
- Location 4: Near the processor fan, microphone in PC casing

As can be seen in Figure 5.9, the microphone location has an important effect on the achievable noise attenuation at the error sensor location. The worst results are obtained when the reference microphone is located in the PC casing, next to the processor fan. In such case, only some tonal control can be achieved. When the reference sensor is also located inside the PC casing, but closer to the power supply block fan, a slightly larger attenuation can be obtained, but the control of broadband noise is still very limited. We believe that this is due to the nature of the broadband noise radiated. Such noise is airflow induced. Most of the turbulences are generated on the ducted side of the fans (flow side). Thus, the reference sensor has to be located in that area to sense the disturbance appropriately.

The microphone located near the power supply block fan, in the duct, yields to the largest achievable control with a coherence above 0.8 over most of the frequency range of interest. The most probable explanation to that phenomenon is that the power supply block fan is the major noise contributor at the duct outlet.

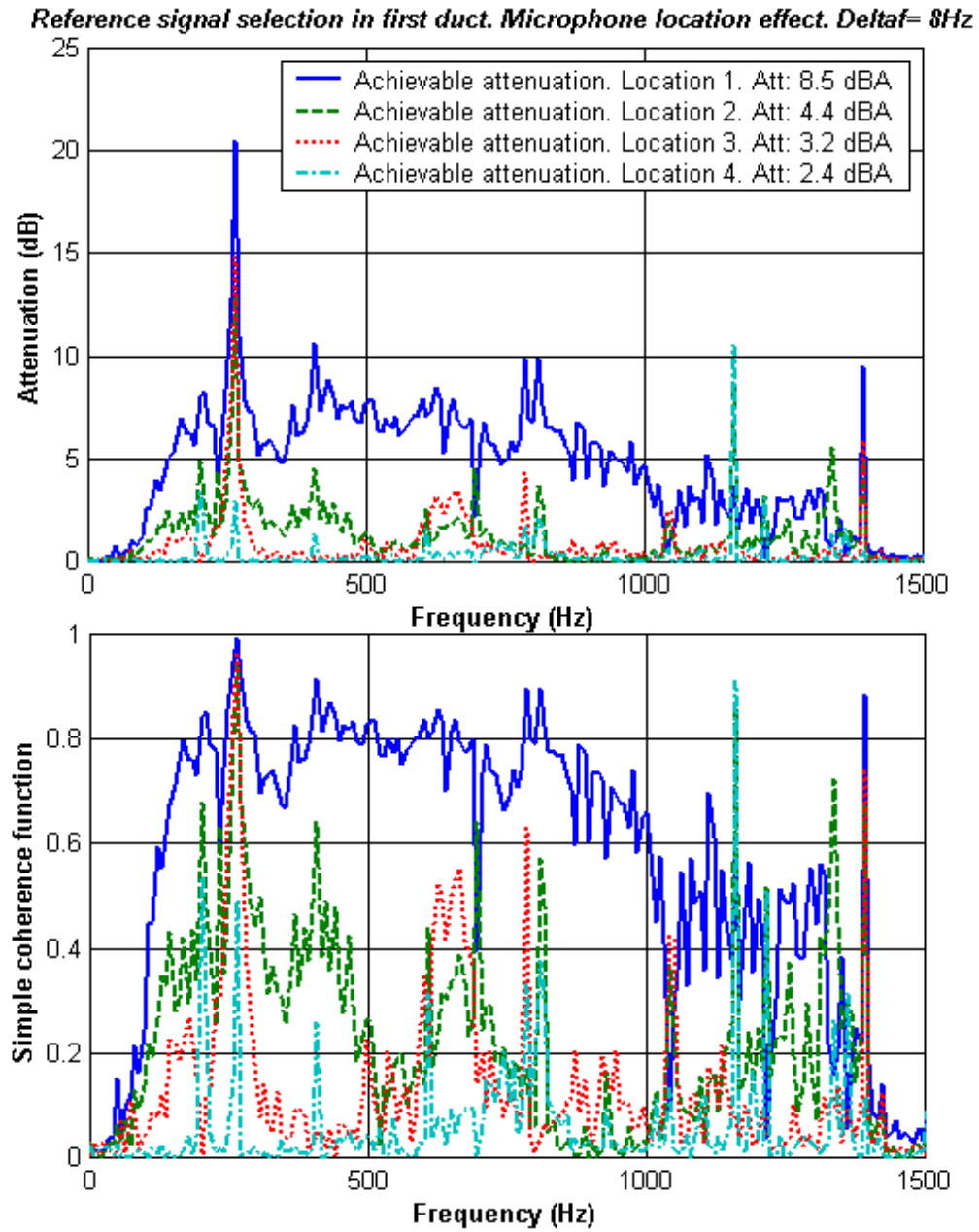


Figure 5.9 Reference microphone location effect on the achievable control in duct.

c. Windscreen Effect

Both the error and reference microphones are operating in a turbulent airflow medium. Here, we want to investigate the effect of the airflow on the measured signal. The microphones are usually not designed to operate properly under such conditions and usually require the use of a windscreen that is acoustically transparent and isolates the microphone diaphragm from the pressure fluctuations of the airflow.

For the following experiments, the achievable noise control at the duct outlet is simulated when both the power supply block and processor fans are operating for two different configurations. In both cases, the error microphone is located at the duct outlet and the reference microphone is placed next to the power supply block fan. The only difference between the two measurements is the presence or not of a windscreen (made out of foam) on the sensors. Results presenting the achievable attenuation at the error sensor and the coherence function are shown in Figure 5.10.

We observe that the windscreen has a significant impact on the coherence between error and reference signals. This is due to the corruption of both signals by measurement noise due to the airflow in the duct. The measurement noise does affect the control of both broadband noise and pure tones. For the future experiments involving pressure sensing in presence of airflow, windscreens will always be used.

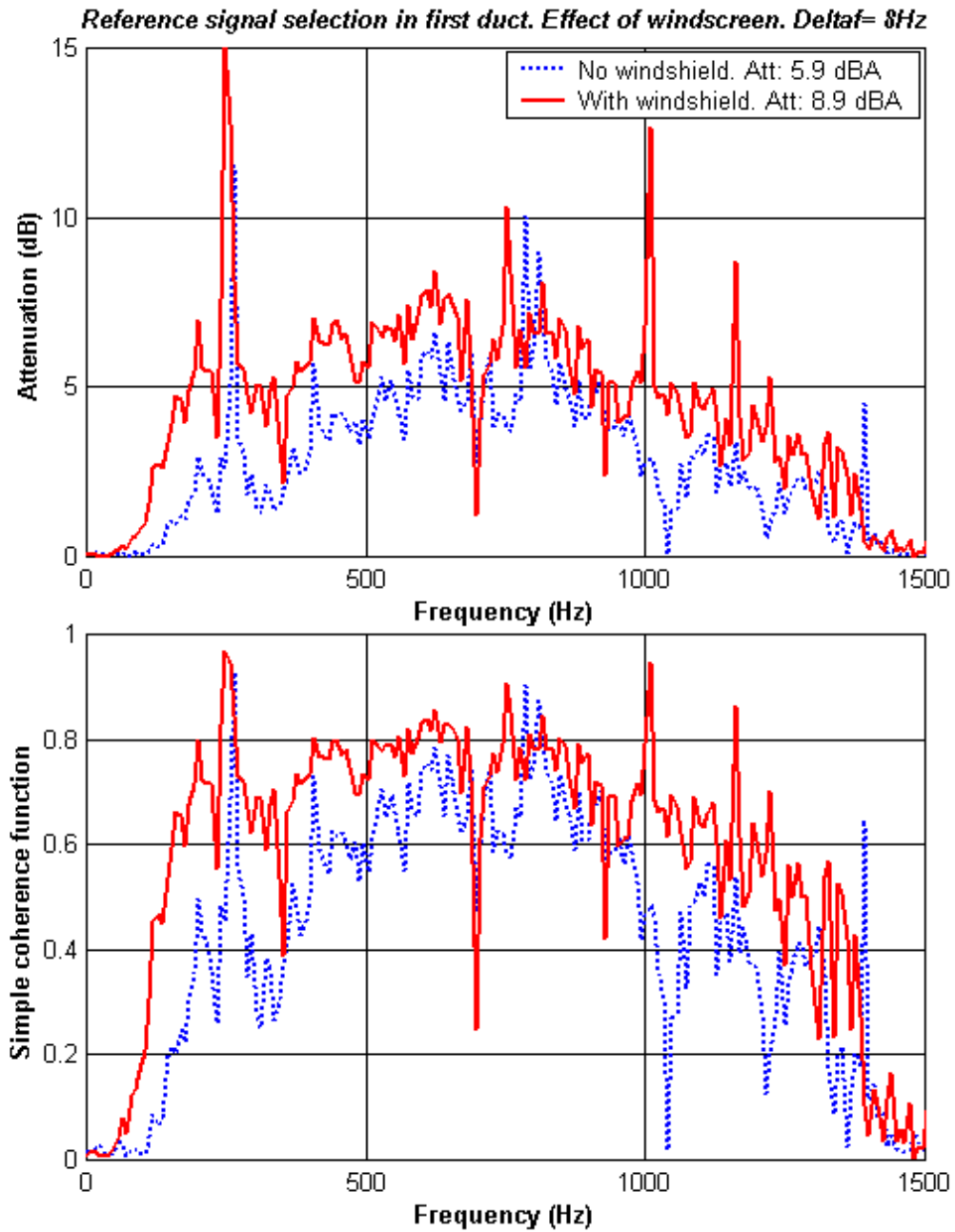


Figure 5.10 Microphone windscreen effect on the achievable control in duct.

d. Sensor Precision Effect

As we have said above, we first used B&K 1/2" high precision condenser microphones as reference and error sensors in the duct before switching to more compact and affordable PCB/Acousticel 1/4" piezo microphones. Since the PCB sensors have a lower signal to noise ratio than the B&K sensors, the controller performance could possibly be reduced when using those PCB units rather than the B&K's.

In the following experiments, the achievable noise control at the duct outlet is compared when using both types of microphones as reference and error sensors. The reference microphone is always located upstream, near the power supply block fan air outlet. Results are shown in Figure 5.11. Contrary to what we expected, it appears that more noise reduction is achievable when using the PCB microphones even though those sensors present a lower signal to noise ratio than the B&K's.

The B&K sensors present better technical characteristics than PCBs such as a very flat and extended frequency response function, a low noise floor and little sensitivity to 60Hz noise. Nevertheless, for this particular application, the noise levels measured are large enough so that noise floor is not an issue. Also, the sensors do not need to have a flat or extended response function (measurements are made up to 1.5KHz only). The results thus demonstrate that the high precision B&K microphones are not needed in this particular problem.

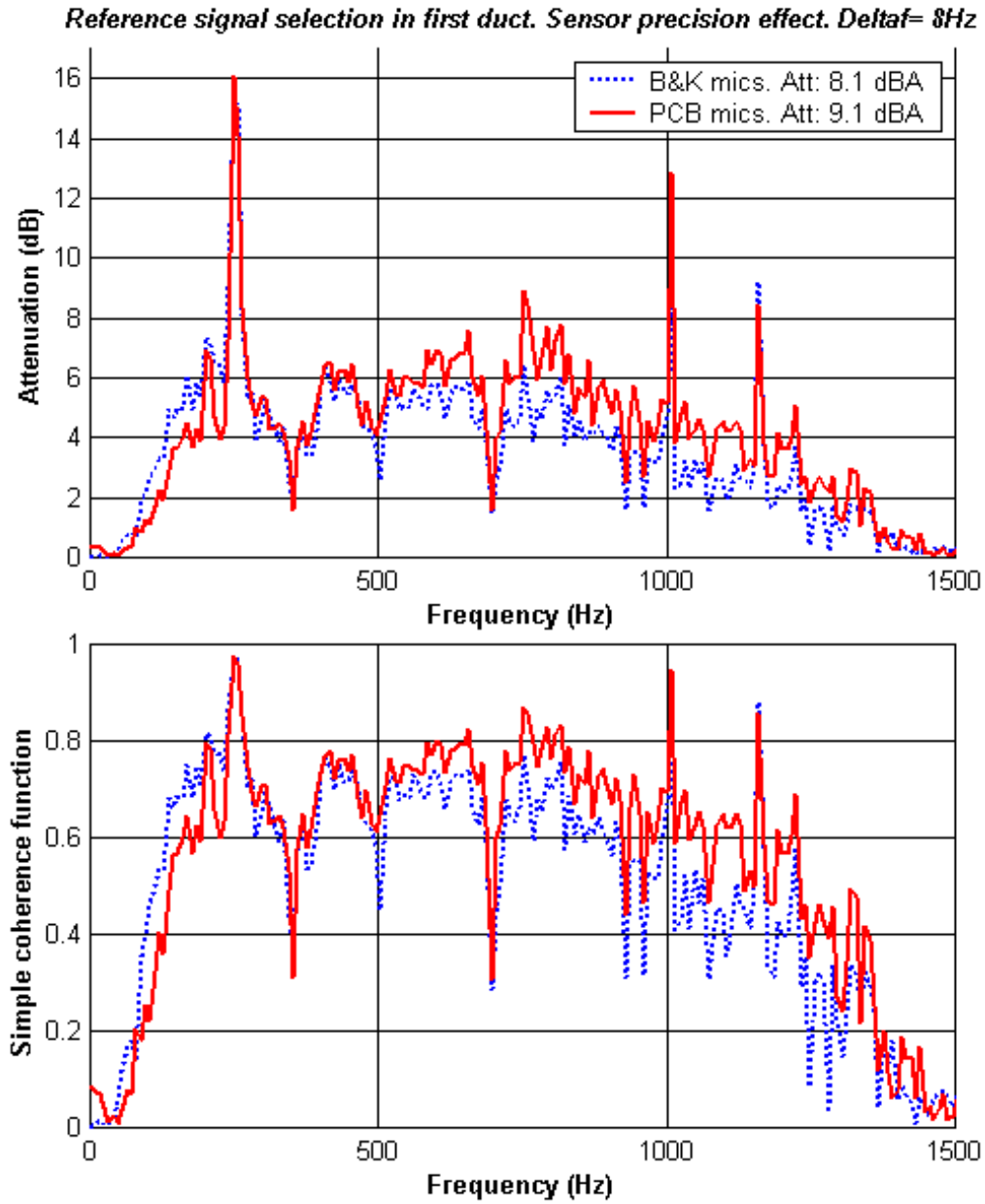


Figure 5.11 Microphone precision effect on the achievable control in duct.

5.3.1.3. Causality Investigation

a. Group Delay in the Analog Filters.

In section 2.1.4, the notion of Group Delay in Linear, Time Invariant systems was introduced. It was stated that the Group Delay is proportional to the slope of the phase response function of the linear system:

$$\delta(j \cdot \omega) = -\frac{d(\vartheta(j \cdot \omega))}{d\omega} = -\frac{d(\vartheta(j \cdot \omega))}{2 \cdot \pi \cdot f} \quad (5.10)$$

We can distinguish two types of LTI systems. The first category corresponds to linear phase systems where the phase response is linearly related to the frequency, which leads to a non-frequency dependent Group Delay:

$$H(j \cdot \omega) = \exp(-j \cdot \omega \cdot t_0) = |H(j \cdot \omega)| \cdot \exp(j \cdot \vartheta(j \cdot \omega)) \quad (5.11)$$

$$|H(j \cdot \omega)| = 1 \quad \vartheta(j \cdot \omega) = -\omega \cdot t_0 \quad (5.12)$$

$$\delta(j \cdot \omega) = -\frac{d(\vartheta(j \cdot \omega))}{d\omega} = t_0 \quad (5.13)$$

As was seen in Chapter 2, non-dispersive propagation problems such as sound in the air or longitudinal waves in beams belong to this first category. On the other hand, most of the physical systems and wave propagation problems are frequency dispersive. In such case, there is no linear relationship between the phase and the frequency over wide frequency ranges, which lead to a frequency dependent Group Delay.

In the control system studied here, the Group Delay is a key parameter since it enables us to determine if the control is causal or not depending on the frequency. Indeed, as we have seen before, in order to have a causal control filter, the overall delay in the control path should be less than the direct path propagation delay, between the reference sensor location and the control actuator. Since we consider plane wave disturbances traveling in the air, this propagation Group Delay is straightforward to estimate. Assuming that the reference microphone and the control speaker are separated by the distance d , the Group

Delay for such non-dispersive wave propagation problem is $\delta_{direct} = d/c_0$, where c_0 is the speed of sound in the air.

On the other hand, the cumulative delay in the control path is not easy to calculate. Most of the linear systems constituting this path have a non-linear phase characteristic, which means the input noise will be delayed in various proportions depending on the frequency. The objective here is to minimize the Group Delay induced by each component of the control path from the reference sensor to the control actuator. We tried below to characterize the Group Delay of the various components of the control path.

We can start with the delay introduced by the low-, high- and/or bandpass analog filters. Ithaco adjustable analog filters were widely used during this project for filtering various signals and are thus now characterized. The filter is fed with a white noise of wide bandwidth (20kHz) and the input and output signals of the filter are simultaneously measured, sampling the data at a rate of 20kHz. Figure 5.12 and Figure 5.13 illustrate the Frequency Response Function and Group Delay of a 2nd order low-pass filter whose cut-off frequency is set at 1000Hz. Figure 5.14 and Figure 5.15 correspond to a 2nd order high-pass analog filter whose cut-on frequency is 200Hz.

In both cases, a Matlab simulated analog filter (Butterworth) gives results very close to the measurements, except at higher frequencies (above 7000Hz) where the anti-aliasing filter of the DAQ system rolls off, thus affecting the signal to noise ratio of the measured signals. Thus, for the following filter characteristics investigation, filters simulated in Matlab replace experimental data in order to save time.

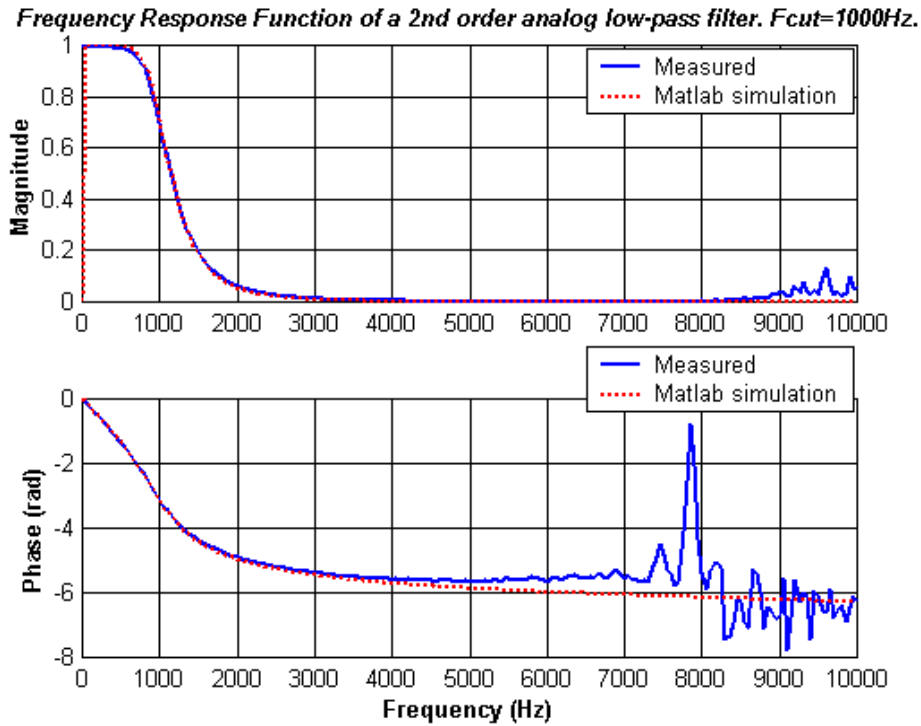


Figure 5.12 FRF of a 2nd order low-pass analog filter.

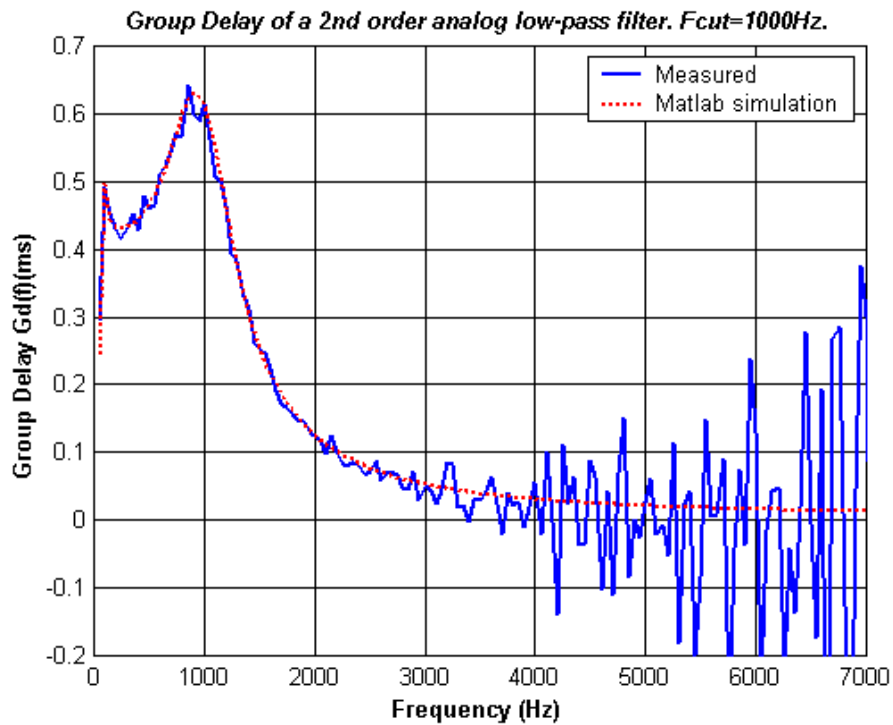


Figure 5.13 Group delay of a 2nd order low-pass analog filter.

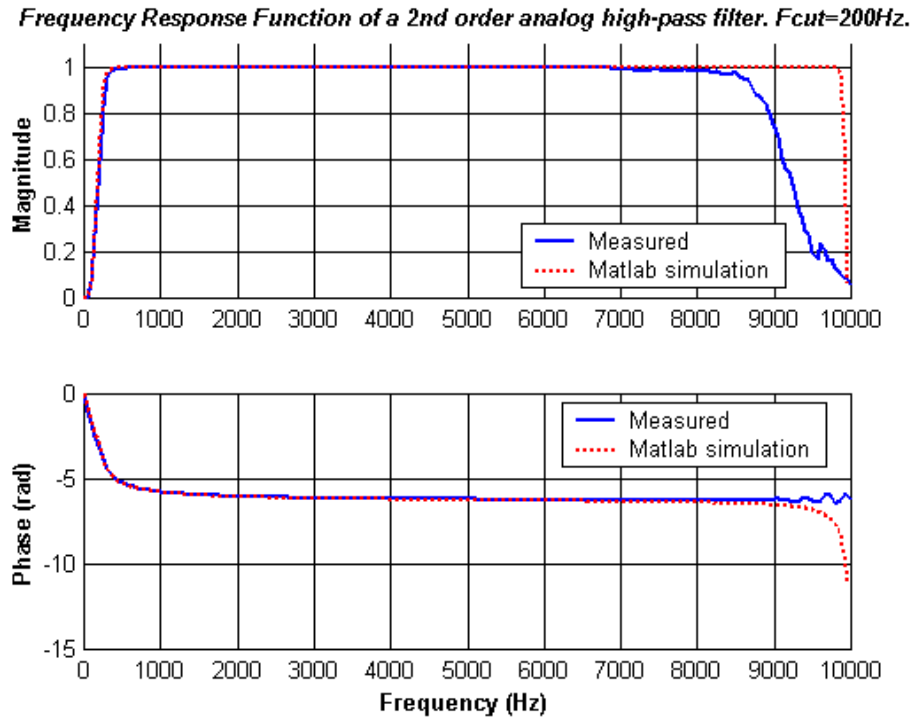


Figure 5.14 FRF of a 2nd order high-pass analog filter.

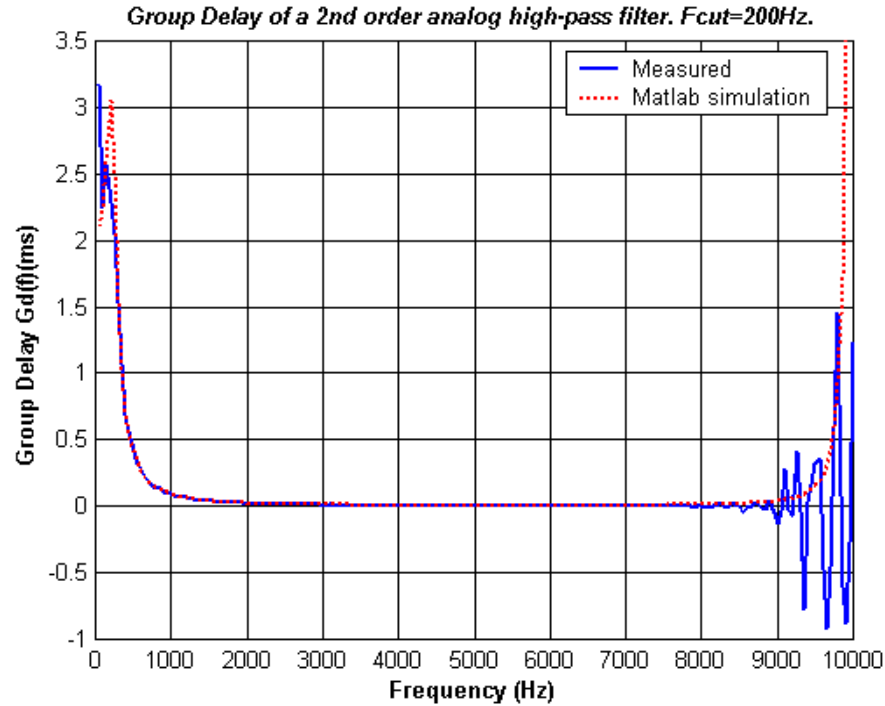


Figure 5.15 Group delay of a 2nd order high-pass analog filter.

From the previous figures, we observe clearly that the filter presents a maximum Group Delay at its cut-off/on frequency band, where the phase response changes rapidly. The higher the filter order, the larger the phase change at the critical frequency and thus the larger the Group Delay obtained in the filter as can be seen in Figure 5.16. For a given cut-on/-off frequency, low-pass and high-pass filters present identical Group Delays provided they are of the same type (Butterworth, Chebychev, etc.) and order.

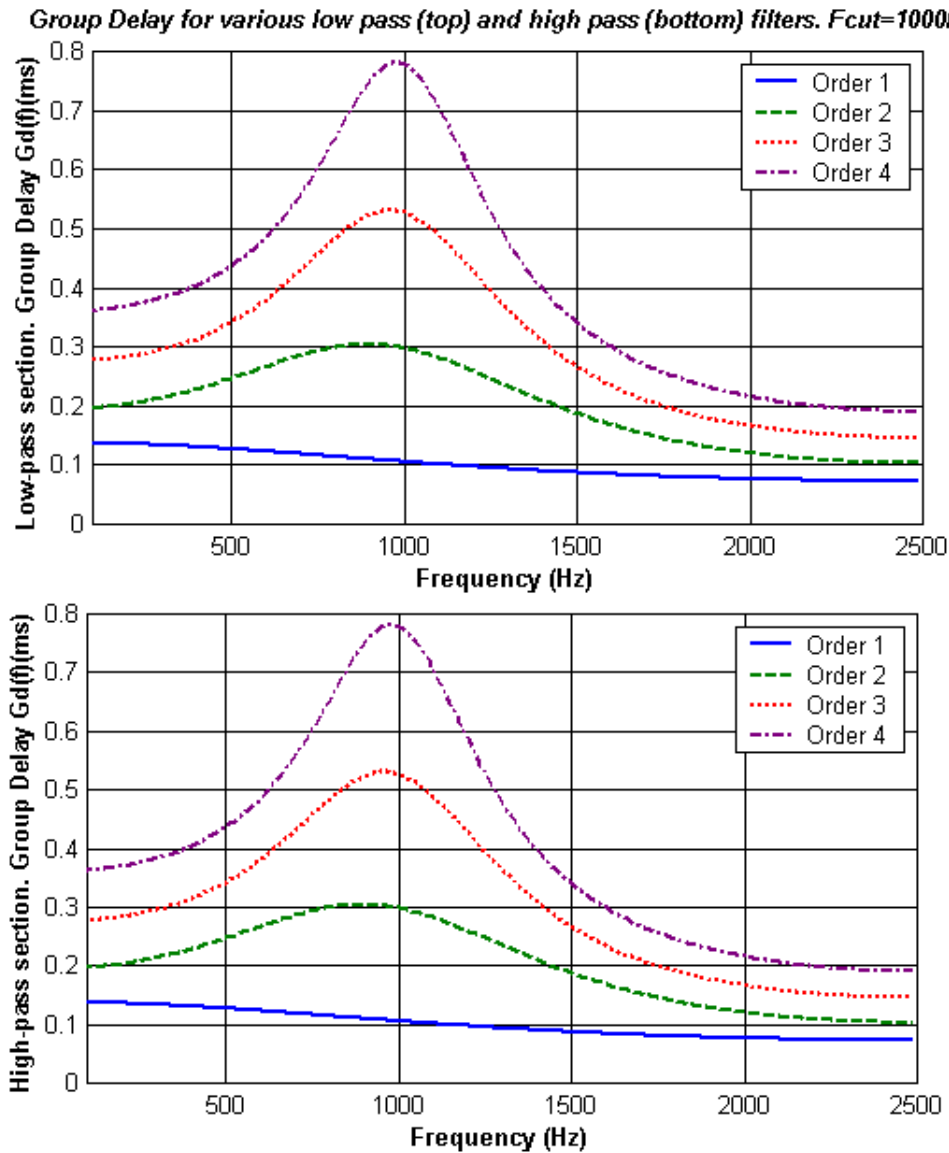


Figure 5.16 Filter order effect on the Group Delay. Butterworth, 1000Hz.

For a given filter order, the phase response is affected by a constant shift (n times $\pi/2$ phase shift with n the filter order), regardless of the filter frequency. On the other hand, the rate of change of the phase and amplitude response depend on the filter cutting frequency. For example, the slope of the amplitude response, in the roll-off region is n times 6dB/octave, with n the filter order. Thus, the lower the cutting frequency, the steeper the slope will be since the octave bandwidth decreases with frequency as can be observed in Figure 5.17. Ultimately, the lower the filter cut-on frequency, the larger the Group Delay for a given filter order and type as can be seen in Figure 5.18.

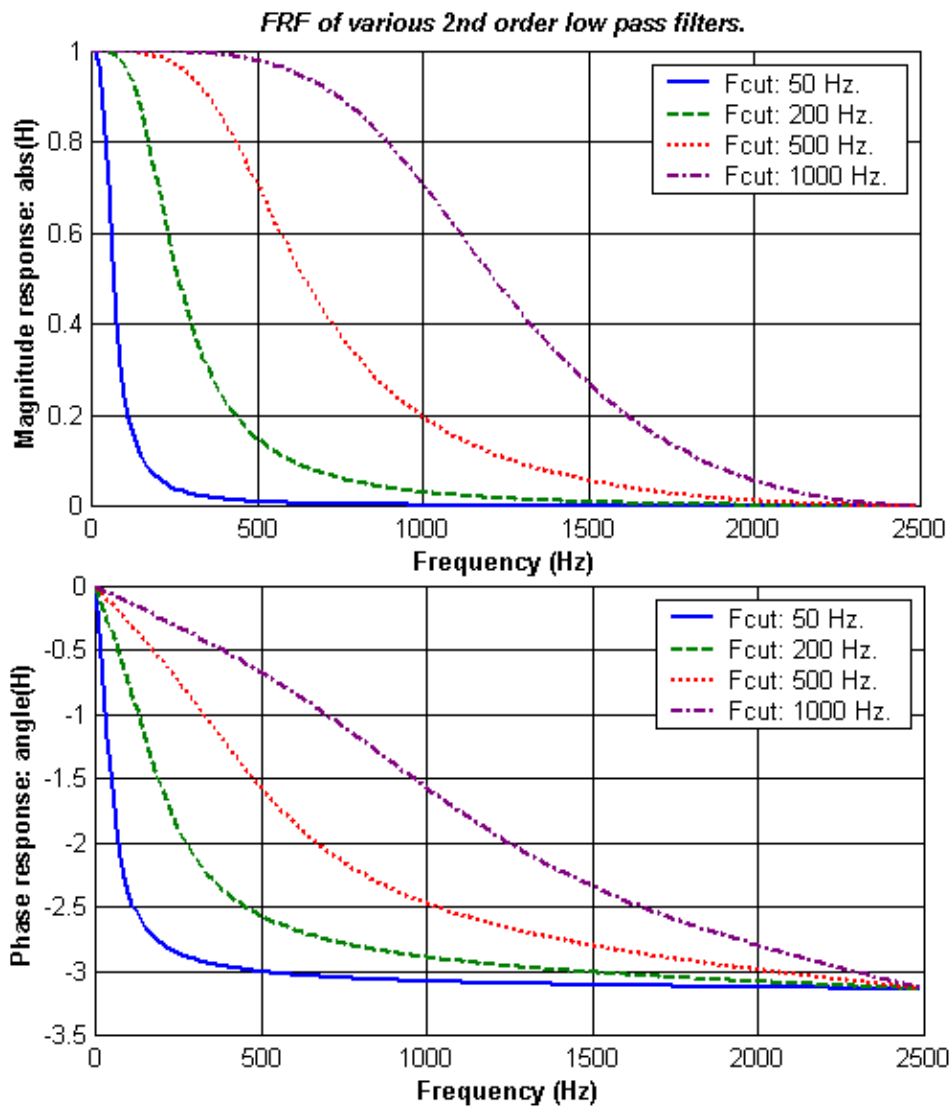


Figure 5.17 FRF of various 2nd order low pass filters, Butterworth type.

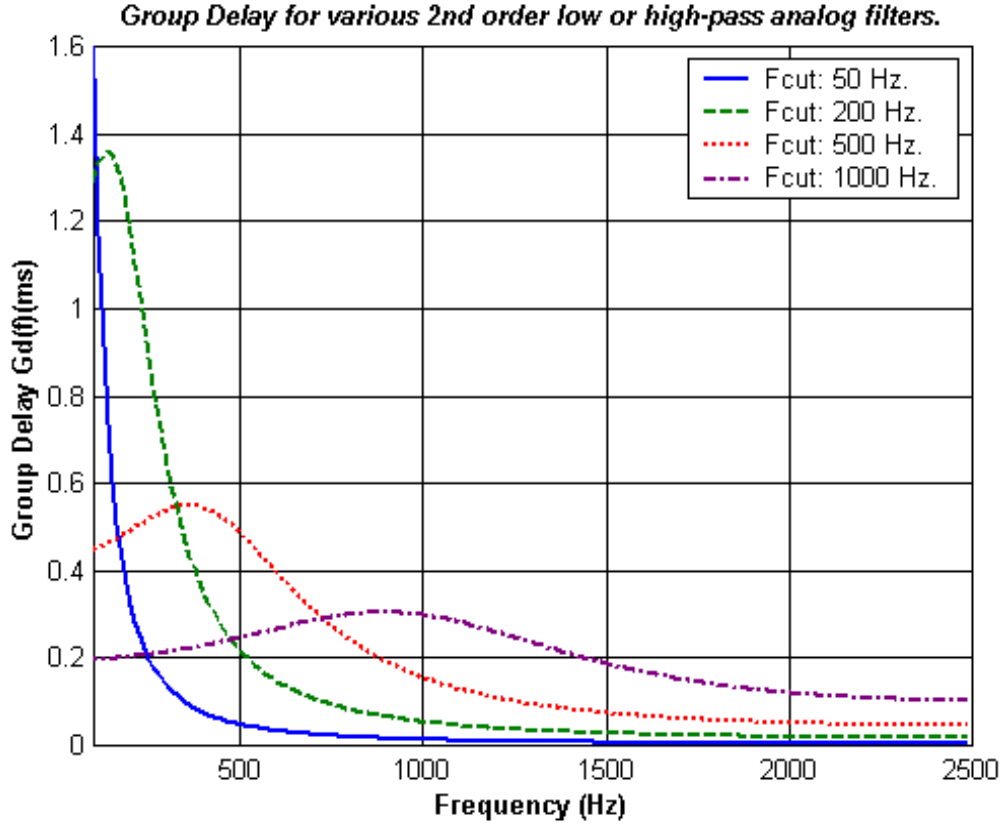


Figure 5.18 Cut-on frequency effect on the Group Delay. Butterworth, 2nd order.

From those observations, we conclude that the use of filters in the control path should be minimized. If filters are needed, we should ensure that they are of low order and that their cut-on/off frequency is well outside the frequency range of ANC. Indeed, we have observed above that the Group Delay is maximum in the transition region and increases with the filter order.

Our early stage noise control systems did not benefit of the above observations based on causality issue. We will illustrate the effect on the controller performance in section 5.3.1.4.b as well as describe the solutions we provided to make the controller operate causally.

b. Group Delay in the Transducers

The transducers represent a second source of delay in the control path. Piezo-electric microphones are used to convert pressure fluctuations into electrical signal. The actuators are of electro-dynamic type. In order to experimentally measure the delay introduced by those transducers, we measure the transfer function in the free field between the electrical signal fed to the speaker and the electrical response of a microphone located at a known distance from the speaker (so that we can compensate for the acoustic transport delay between the actuator and the sensor).

Then, the estimated Group Delay is the sum of the delays introduced by both transducers. Nevertheless, we can reasonably assume that the major contributor to the delay will be the actuator. Indeed, due to the cone mass, the settling time of the electro-dynamic actuator is likely much higher than that of the small piezoelectric transducer. We will not study in detail the dynamics of electro-dynamic actuators. But, the response of such a transducer at lower frequencies can be viewed as a high-pass filter whose order depends on the acoustical load of the driver (free field, sealed or ported enclosure). Thus, it will present very similar characteristics to what has been observed earlier with the analog electrical filters.

Most of the comparisons made below are for Transfer Function measurements with a microphone located in the near field of the speaker. We first want to make sure that the near field measurements lead to similar results as in the far field by comparing the Transfer Function and Group Delays for two separation distances between the source and the microphone (10 and 50cm). The results presented in Figure 5.19 show that even though the near field and far field response are slightly different, the estimated Group Delay is not significantly affected.

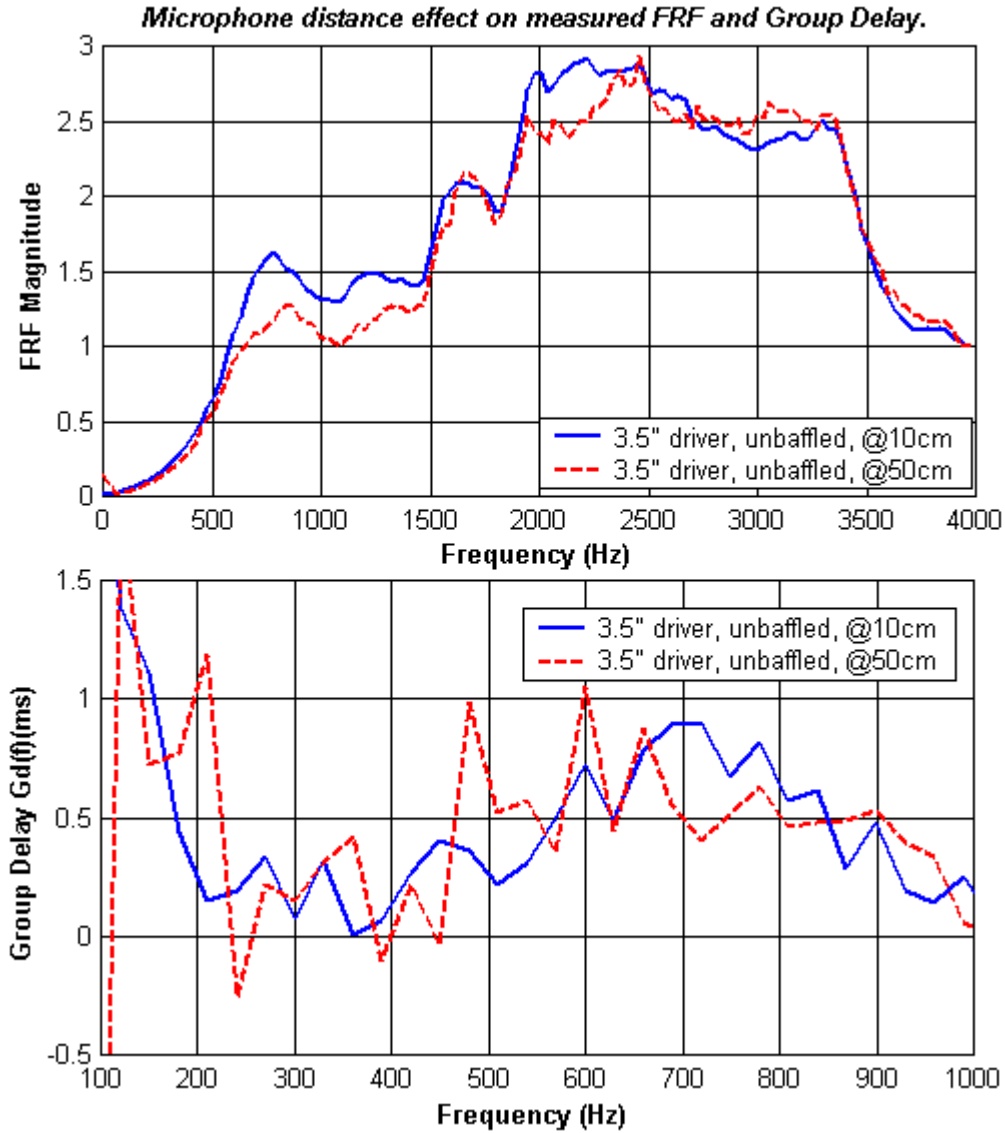
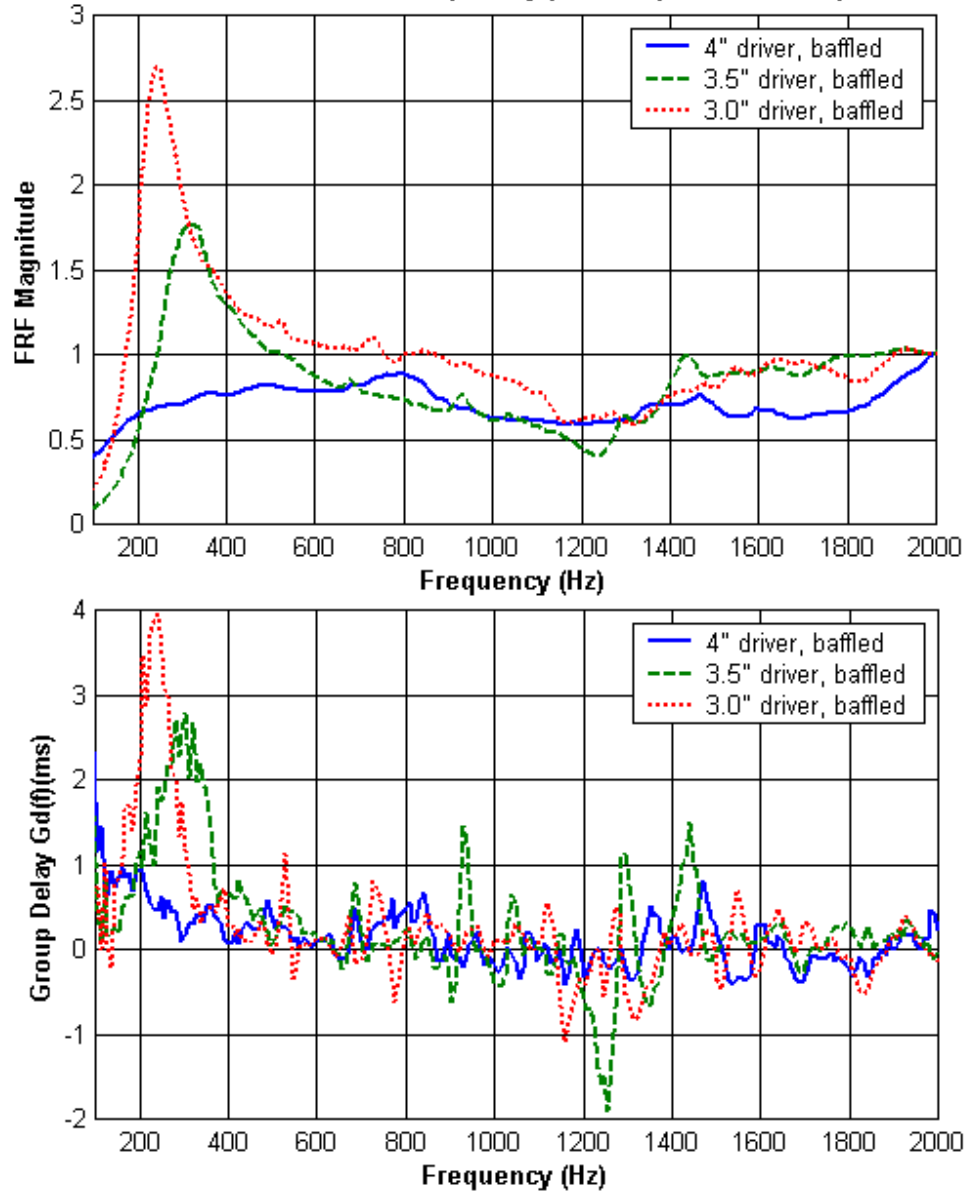


Figure 5.19 Effect of microphone-speaker distance on the FRF and Group Delay.

The next test performed was to study the effect of the driver size on the measured Group Delay. As can be observed in Figure 5.20, the smaller the driver, the larger the Group Delay. Nevertheless, we believe that, rather than the size, the Quality Factor of the resonant mechanical system constituted by the diaphragm and suspension is responsible for the Group Delay [28]. Indeed, when the suspension of the mechanical system is not very stiff and the damping low, then the quality factor is large, which in turns leads to a large Group Delay in that frequency range.

Effect of driver size on FRF and Group Delay (baffled speakers, microphone at 10cm)**Figure 5.20 Effect of the driver size on the FRF and Group Delay (baffled drivers).**

The final test performed was intended to check the effect of the acoustical load on the driver response and Group Delay. Results are shown in Figure 5.21 and Figure 5.22. As can be observed in Figure 5.21, the un-baffled driver yields generally to the lowest Group Delay over a wide frequency range, at the expense of a poor authority at lower frequencies (where the wavelength is much larger than the diaphragm dimension).

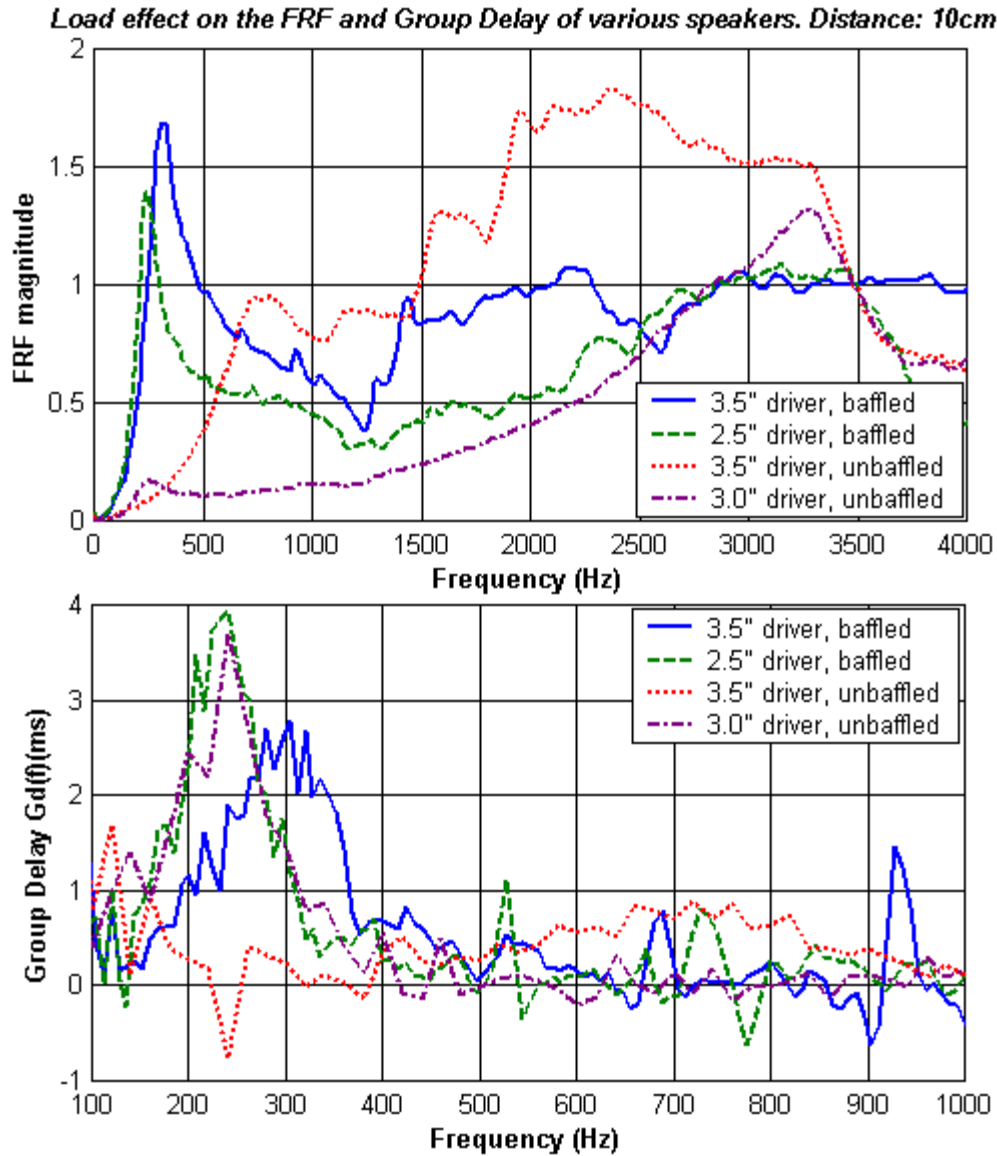


Figure 5.21 Effect of the acoustical load on the FRF and Group Delay.

In Figure 5.22, we see that the driver, in a sealed enclosure, has the highest authority at low frequencies (compared to the baffled configuration), at the expense again of the Group Delay. We opted for the sealed enclosure load for all our designs since the secondary source strength was a priority. Nevertheless, the Group Delays observed here are much larger than for the analog filters, often above 0.5ms for frequencies below 500Hz. Thus, the delay introduced by the electro-acoustic transducer is the limiting factor for the ANC performance due to causality issues. For example, a delay of 0.5ms in the

control path translates into a minimal separation distance of 17cm between the reference microphone and the actuator. If we add the delay induced by the digital controller and the analog filters, the control system may not operate causally at low frequencies (below 300 to 500Hz, where the speaker Group Delay becomes higher) since the reference microphone and secondary noise source are not more than 20cm apart in the first duct prototype.

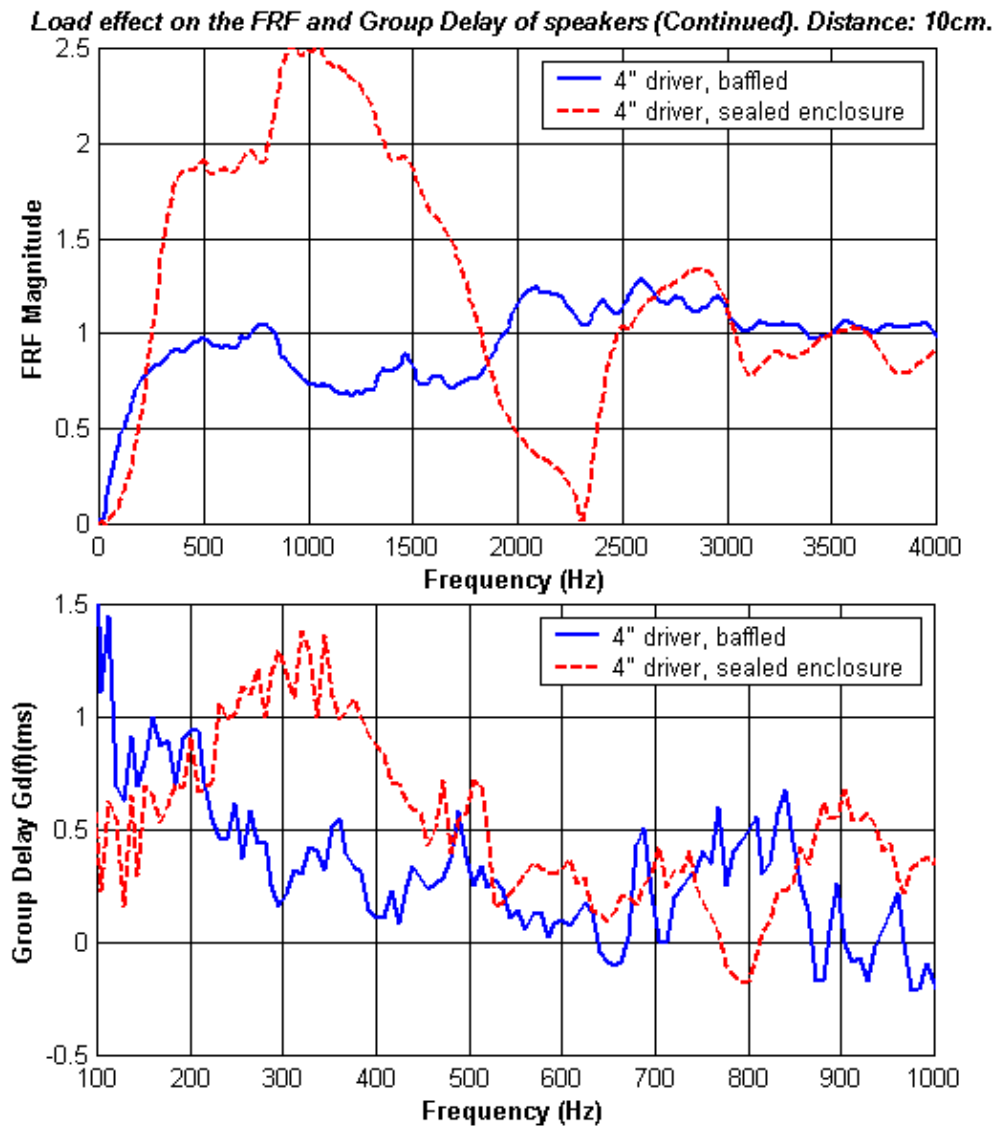


Figure 5.22 Effect of the acoustical load on the FRF and Group Delay (continued).

c. Group Delay in the Controller

The digital controller is the last component in the control path that can add delay. The response of the controller is composed of the responses of the analog to digital and digital to analog converters as well as the adaptive digital filter itself. In our case, the controller used also includes built in anti-aliasing adjustable filters that are present on the input board of the controller.

If the control is causal, the delay in the digital control filter should equal the propagation delay between reference sensor and control speaker diminished of the delays in the control path (actuators, analog filters, A/D and D/A converters, processing delay).

Unfortunately, the delay introduced by the controller could not be accurately measured. However, we can estimate the delay introduced by the converters and the filtering process. Such delay should not be frequency dependent and related to the controller sample rate. Indeed, we can assume that each of the converters delays the signal by 1 sample (or more), which corresponds to $1/F_{\text{samp}}$ seconds with F_{samp} the controller sample rate. Also, the input and output sequences of the digital control filter are delayed by at least one sample due to processing delay. Thus, the minimal lag introduced by the digital controller, excluding any anti-aliasing or reconstruction filter is at least three taps, which corresponds to a duration of $3/F_{\text{samp}}$ seconds.

From this observation, it is clear that our objective is to maximize the controller sample rate in order to minimize the delay. The sample rate has to be at least twice the maximum frequency we are attempting to control. Since our goal is to actively control noise up to about 1200Hz, the controller sample rate should be 2500Hz or higher. At a sample rate of 2500Hz, the delay in the digital controller will be 1.2ms, which translates into a minimum separation of 40cm between the reference microphone and secondary source, assuming there is no delay induced in the other components of the control path. This will definitely lead to an acausal control. On the other hand, a sample rate of 7500Hz induces a delay of 0.4ms, which decreases the minimum distance to 14cm.

5.3.1.4. ANC System Performance

a. Presentation of the Controller

The DSP board used for the ANC experiments on the first duct prototype is a Texas Instruments TMS320C30, interfaced via a host PC. This board is used to implement a single input / single output feedforward digital controller using FIR filters. The control filter update is generated based on the Filtered-X LMS algorithm. The PC interface to the DSP board was developed in the VAL Labs using the Labview software.

There are two major benefits to using such a graphical interface. First, it allows us to easily modify the settings of the control system. The interface automatically translates the settings chosen by the user into a code (typically in assembly language) that is then uploaded to the DSP board. Many control parameters such as the sample rate, the number of taps in the control and secondary path filters are adjustable via the Labview interface. Where as the previous parameters are set prior to control, the rate of convergence factor μ for the LMS update algorithm can be adjusted in real time.

The second major benefit of the graphical interface is that it shows the evolution of the different signals and filters as the controller converges. We can thus observe the control filter shape (impulse response or FRF) and diagnose, for example, an acausal control (maximal filter value at the tap 0) or a lack of taps (the impulse response does not fit completely in the filter block).

We can also observe the reference, control and error signals in real time as well, which is very useful to diagnose saturation of the controller for example. An important feature is the possibility to look at the evolution of the rms level of the error signal relative to the level prior to control. This gives a real time estimate of the evolution of the achieved attenuation that can then be used to interactively modify the convergence factor if needed. For example, once the attenuation stabilizes, the controller has converged. It is then advisable to minimize the convergence factor in order to keep the controller stable and have it perform optimally (assuming the primary path is not varying largely, which is true in our case).

b. Causality Effect on the Controller Performance: Power Supply Block Fan Noise.

In order to illustrate the effect of causality problems on the controller performance, we ran three ANC experiments with the first duct prototype attached to the PC casing, and the power supply block fan as primary noise source. For all those experiments, the control architecture was identical, with a PCB/Acousticel microphone near the power supply block fan air outlet as upstream reference sensor, the 2.5" driver, characterized earlier, in a sealed enclosure as secondary noise source, and a PCB/Acousticel microphone at the duct outlet as error sensor. The block diagram of the control system is shown in Figure 5.23. The control settings varying among the three tests were the analog filters (R, C and E) characteristics and the controller sample rate (c.f. Table 5.1). We could thus progressively increase the overall delay in the control path so that the controller performs with increased causality problems. The control and identification filters were 80 taps long for all the tests, which was sufficient to model the control and secondary paths accurately at the different sample rates used. The controller incorporates a feedback removal option (see section 2.3.2.1.e) that was nevertheless not needed for the duct experiments. Indeed, the reference microphone was not sensitive to the secondary source signal propagating upstream the duct.

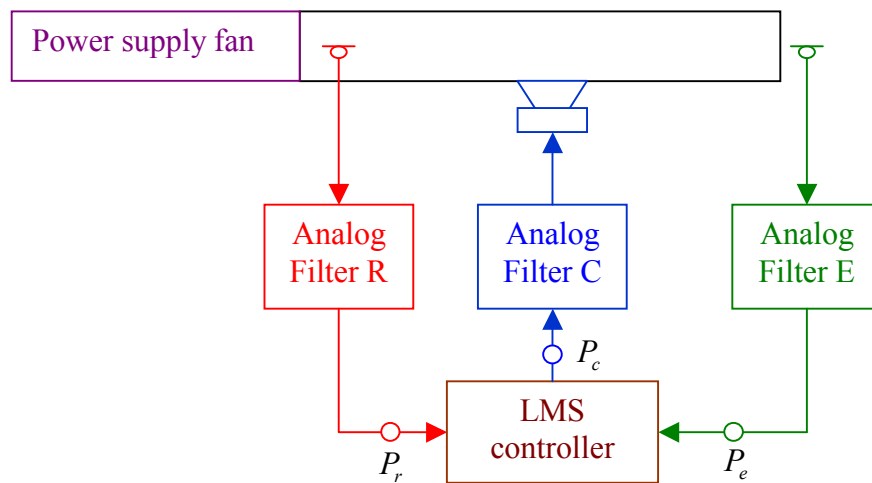
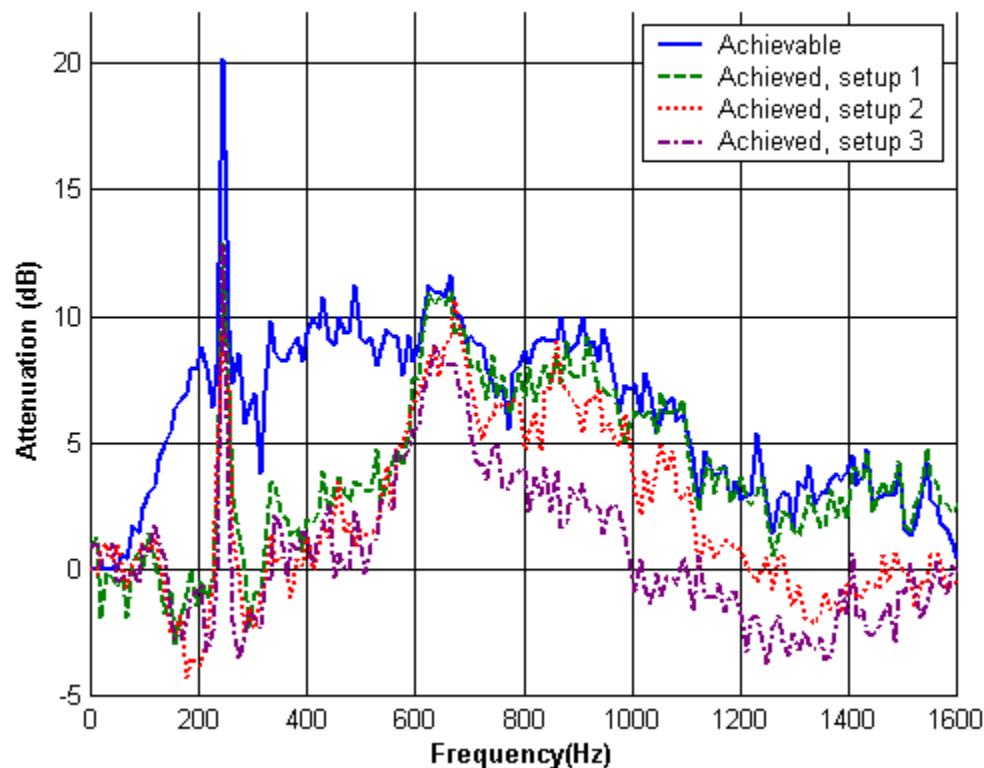


Figure 5.23 Noise controller block diagram, causality tests.

<i>Control setup settings</i>			
<i>Settings</i>	<i>Setup 1</i>	<i>Setup 2</i>	<i>Setup 3</i>
Filter R:	Highpass 200Hz	Bandpass 200-1600Hz	Bandpass 200-1600Hz
Filter C:	Bandpass 20-3150Hz	Bandpass 20-1600Hz	Bandpass 20-1600Hz
Filter E:	Bandpass 20-1600Hz	Bandpass 20-1600Hz	Bandpass 20-1600Hz
Controller Fsamp:	8000Hz	8000Hz	4000Hz

Table 5.1 Control setup settings.

The attenuation achieved at the duct outlet for the different tests is compared to the theoretically maximum attenuation (based on the coherence between reference and error sensors) in Figure 5.24. The theoretically achievable attenuation simulates a control that is causal with a fully authoritative secondary noise source and no external noise present.

Causality effect on controller performance. First duct prototype. Power Supply fan noise**Figure 5.24** Causality effect on the ANC performance in the first duct prototype.

The first observation concerns the control of pure tones (power supply block fan BPF at 250Hz in the present case). Regardless of the controller sample rate or Group Delay in the control path, we achieve the same control of the BPF. This illustrates the fact that the causality is not an issue for purely predictable noises. The achieved control (12dB) is less than predicted (20dB). This could be due to limitations in the secondary noise source authority.

The second observation concerns the broadband noise control. In the setup with the least delay in the control path (setup 1), the achieved control reaches the theoretical limit for frequencies above 600Hz. There are two possible reasons for the lack of control below 600Hz. First, as we have seen before, the secondary noise source induces a large delay at lower frequencies. Also, the reference signal is high-pass filtered at 200Hz to prevent the controller from saturating due to large measured signals at low frequencies (flow noise). This filter introduces a delay in the cut-off region as seen earlier. Thus, the causality issue is most probably a limiting factor for control below 600Hz. The second reason might be a lack in control authority due to the poor performance of the secondary noise source at lower frequencies.

When the delay in the control path is increased, due to both the addition of filters whose cut-off frequencies are in the noise control range as well as to the reduction of the controller sample rate, the controller performance is reduced accordingly. Thus, the performance of the ANC system is directly related to the causality problems.

We observe here that the optimal setup (setup 1) had no low-pass filter for the reference signal. The reason is that the DSP board incorporates anti-aliasing filters that are automatically set based on the controller sample rate. On the other hand, the control filter incorporates a low-pass filter section because our controller did not integrate a reconstruction filter. By over-sampling the controller, we could set that reconstruction filter frequency to a value well beyond the control range so that the induced delay was minimal. Also, the anti-aliasing filter, whose cut-off frequency is set to approximately half the sample rate frequency, would not introduce much delay for the same reason.

c. In Duct Performance for the Fan Noise Control.

Based on the results obtained with the previous test, we performed control tests while both the power supply and processor fans were operating. The processor fan was nevertheless structurally decoupled from the PC casing for the following experiments.

The control setup is identical to the previous experiments. The reference signal is high-pass filtered at 200Hz for the same reason as before. The controller is set to a sample rate of 8000Hz, which fixes the cut-off frequency of the anti-aliasing filters to approximately 4000Hz, a value well beyond the maximal frequency of control. The error signal is band-pass filtered between 20 and 1600Hz to limit the frequency range of Active Noise Control to plane wave propagation. The control signal is band-pass filtered between 20 and 3150Hz. The low-pass section is used to filter out the reconstruction noise induced by the digital to analog converter. The high-pass section prevents the controller from saturating the secondary noise source. As before, 80 taps are used for the identification and control filters. Those controller settings are shown in Figure 5.25 and the control results are presented in Figure 5.26 and Figure 5.27.

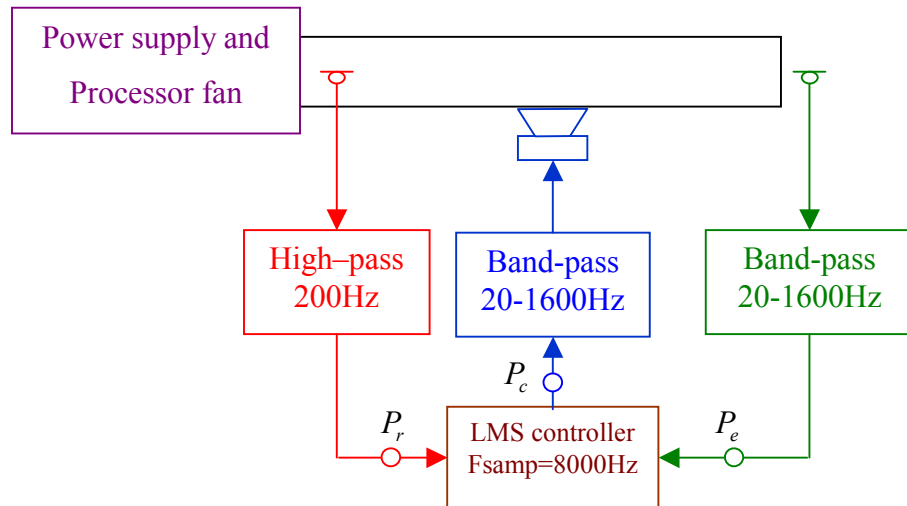


Figure 5.25 Noise controller block diagram, final setup for the first duct prototype.

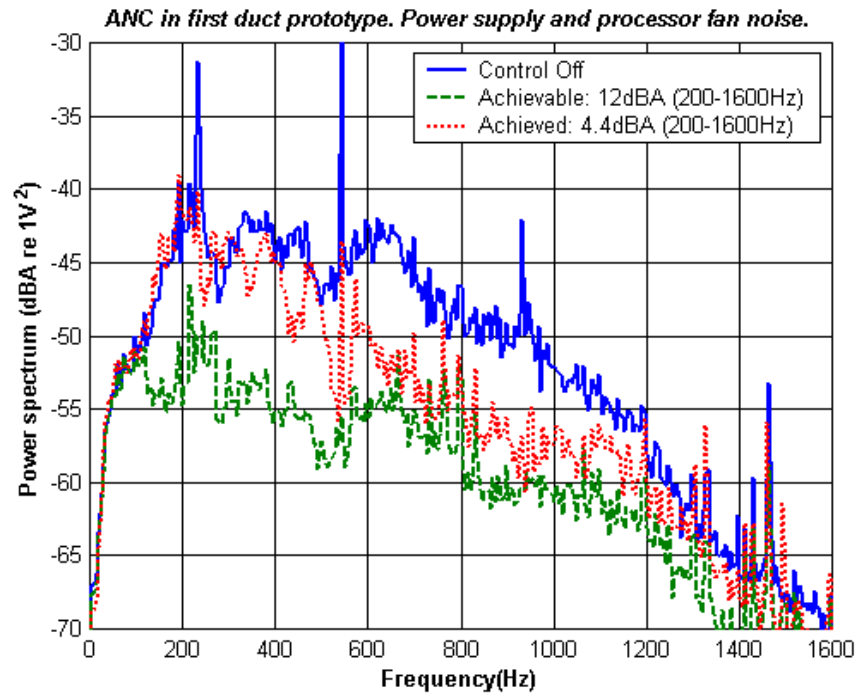


Figure 5.26 Fans ANC in first duct prototype. Error signal power spectrum.

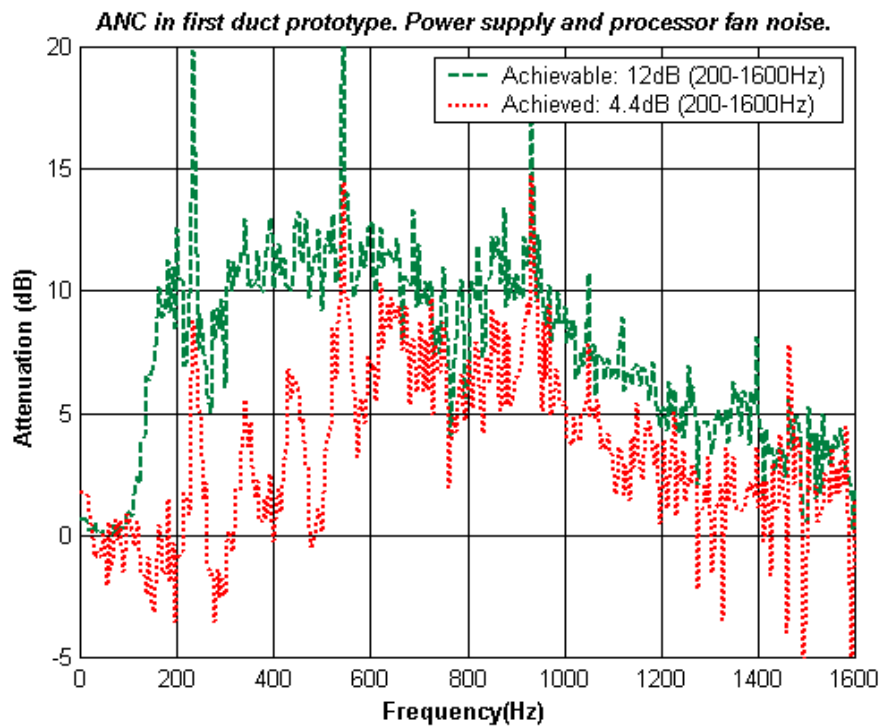


Figure 5.27 Fans ANC in first duct prototype. Error signal attenuation.

As has been observed earlier, no control is achieved below 600Hz. Unfortunately, the control settings could not be further optimized since the C30 controller would not allow us to sample at a higher rate than 8000Hz using 80 taps per filter. The other solution to minimize the causality problems would be to increase the separation distance between reference microphone and control source, which could not be done with the first duct prototype.

As we have explained in section 4.2.2.3.b, the results obtained above lead to the design of a new duct prototype using better performance secondary noise sources and maximizing the distance between reference sensor and control speaker. Moreover, the new prototype will aim at controlling both the PC air inlet and outlet noise. Thus, provided these airborne noise radiation paths are dominant components of the PC radiated noise, a global control of PC noise should be achieved. We have seen in Chapter 3 that the structure borne noise radiation paths have been well controlled passively. But, we will observe that the PC casing, even after the addition of inlet and outlet ducts will present a low noise transmission loss which will limit the global performance of our global noise control system.

5.3.2. Second Duct Prototype

5.3.2.1. Control System Presentation

The new prototype is composed of two separate ducts. The processor fan has been relocated to the front of the PC, at one end of the inlet duct, blowing air into the PC. The outlet duct did not change location but is now aimed at controlling the power supply block fan noise mainly. Pictures of the new ducts are shown in Figure 5.28.

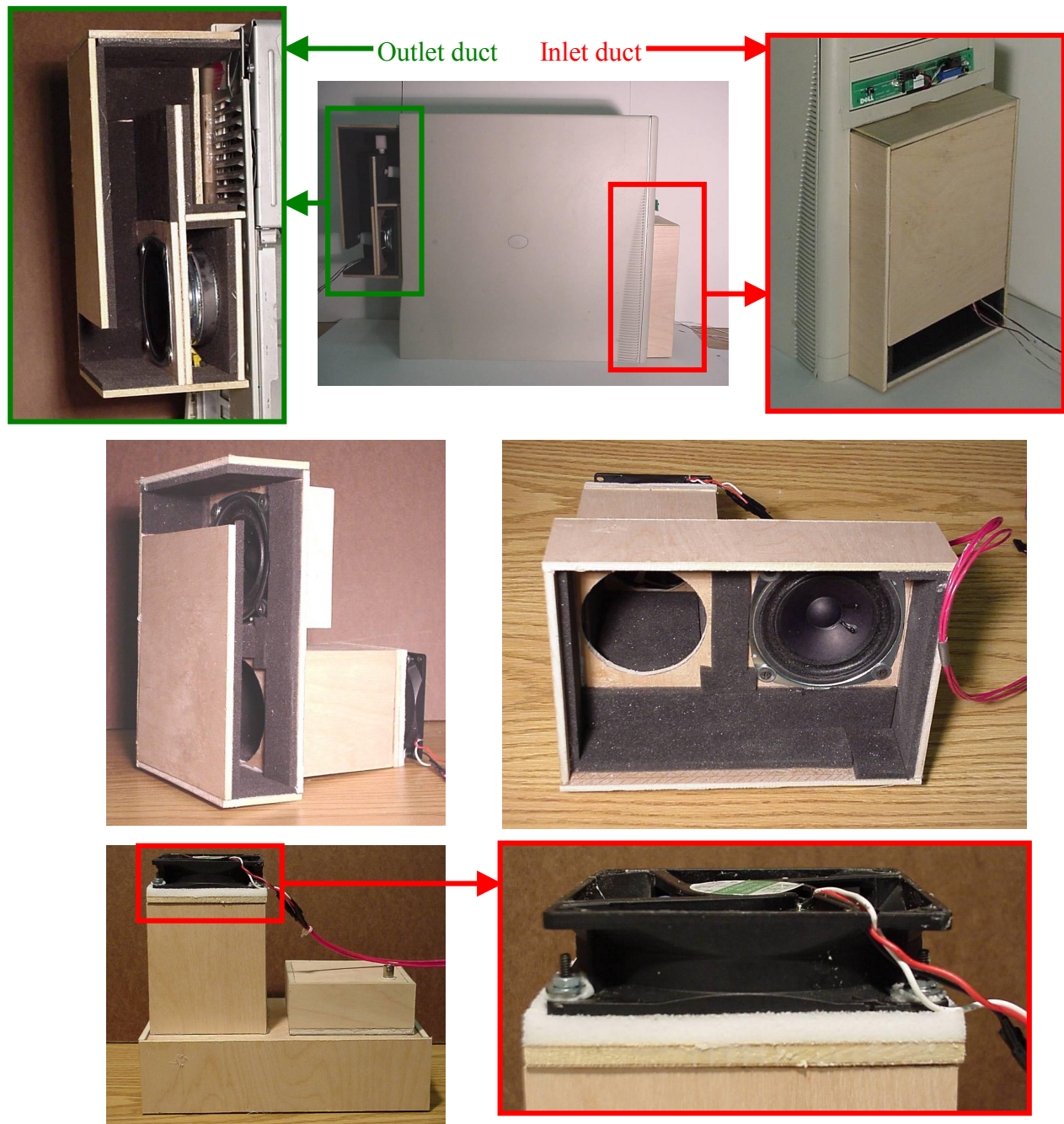


Figure 5.28 Pictures of the second duct prototype.

The new ducts have a slightly different geometry from the first prototype. The internal ducts cross-section is wider and the secondary noise source used is now an audio 4" driver with better performance characteristics than the original driver as we have seen in section 4.2.3.1. We observe in Figure 5.28 that the control source is now located at the outlet for both ducts. Thus, the error microphone will be located in the near field of the secondary noise source. This is usually not recommended in practical designs since the controller tends then to control the near field noise (with both plane and higher order waves) radiated from the secondary noise source rather than the propagating primary disturbance. Nevertheless, we chose that option in order to maximize the delay between reference sensor and control speaker as well as minimize the delay in the secondary path. Indeed, the latter is a source of controller instability (c.f. explanations in section 0) and leads to large identification filters when the controller sample rate is high due to the propagation lag between control source and error sensor.

The reference and error sensors used were of the same type as in the original duct prototype. These were PCB/Acousticel 1/4" piezo-type microphones. We would also use the signal from the tachometer built in the fans as well as accelerometers mounted on the vibrating sources as reference signals. The pressure sensors are always equipped with a windscreen since we have seen in the previous section that they were sensitive to airflow induced measurement noise. The reference pressure sensor was always located in the near field of the primary noise source. This leads to a separation distance between the reference sensor and control source of approximately 30 and 23 cm for the inlet and outlet ducts respectively. As compared to the first duct prototype, the new separation distance between reference sensor and control actuator is larger, which may help to improve the controller performance since more delay in the control path is tolerated.

Since there are now two ANC systems (one per duct), we will first use a C30 board with a SISO controller and then a C40 board that allows MIMO control. Otherwise, the control architecture will be identical to what has been developed for the first duct prototype.

5.3.2.2. Reference Signal Selection

We performed the same experiments as for the first duct prototype. One accelerometer is mounted on the processor fan casing. The other one is attached to the power supply block. The reference microphones are located in the duct, nearby the primary noise sources. We assume a single input / single output controller. Thus, by using the simple coherence function between the reference and error signals, we can predict the most achievable attenuation. The results at the error sensor location are shown in Figure 5.29 and Figure 5.30 for the inlet and outlet ducts respectively.

As was observed in section 5.3.1.2.a, the microphone is the only reference sensor coherent with both the broadband and pure tone noise contribution at the error sensor location. The accelerometer or the tachometer, if used as unique reference sensor, will lead to the exclusive control of the pure tones radiated by the fans (BPF and harmonics), the accelerometer giving slightly better results than the tachometer. Nevertheless, we observe in the inlet duct that a tone at 250Hz is not controlled by either of these sensors. This is because this tone is actually generated by the power supply block fan (BPF). It can only be controlled using the reference microphone or an extra reference signal coming from the power supply section (accelerometer or tachometer). In practice, we will be using a fully coupled two input / two output controller. Thus, the reference signals coming from each primary noise source will be used to control the noise at each duct outlet. In such case, assuming accelerometers are the only reference sensors used, the power supply block fan tones propagating in the inlet duct will be controlled as well. In order to simulate such control, we should use the multiple coherence function rather than the single coherence function as was developed in section 2.1.3.

Overall, looking at the results in Figure 5.29 and Figure 5.30, the control system looks promising with an overall predicted attenuation at the error microphone of 9 and 11.5dBA for the inlet and outlet ducts respectively, between 200 and 1200Hz. This demonstrates that the reference signals (the microphone in particular) sense the primary disturbance appropriately and only small measurement noise is present in the signals.

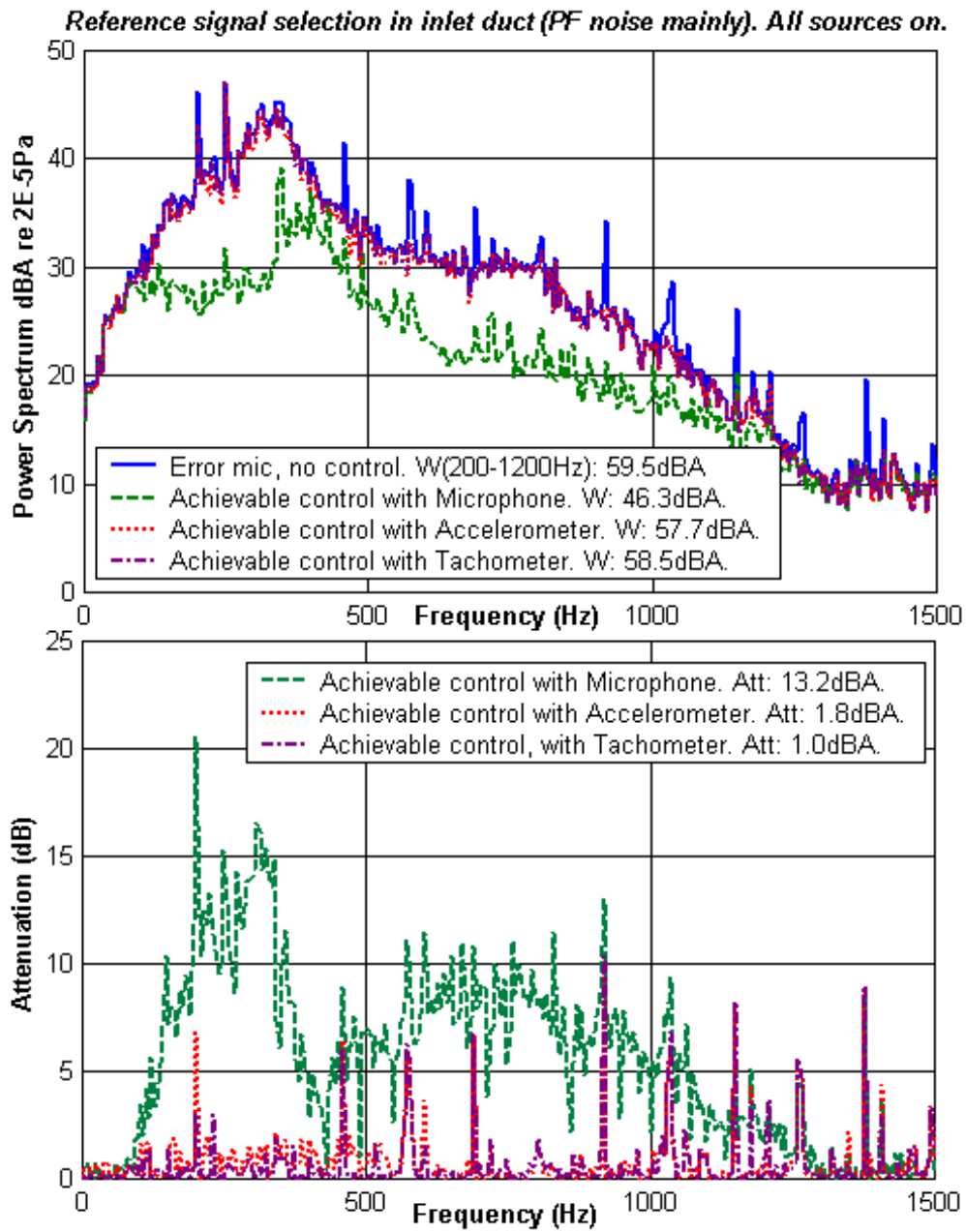


Figure 5.29 Single channel ANC in inlet duct. Reference signal selection.

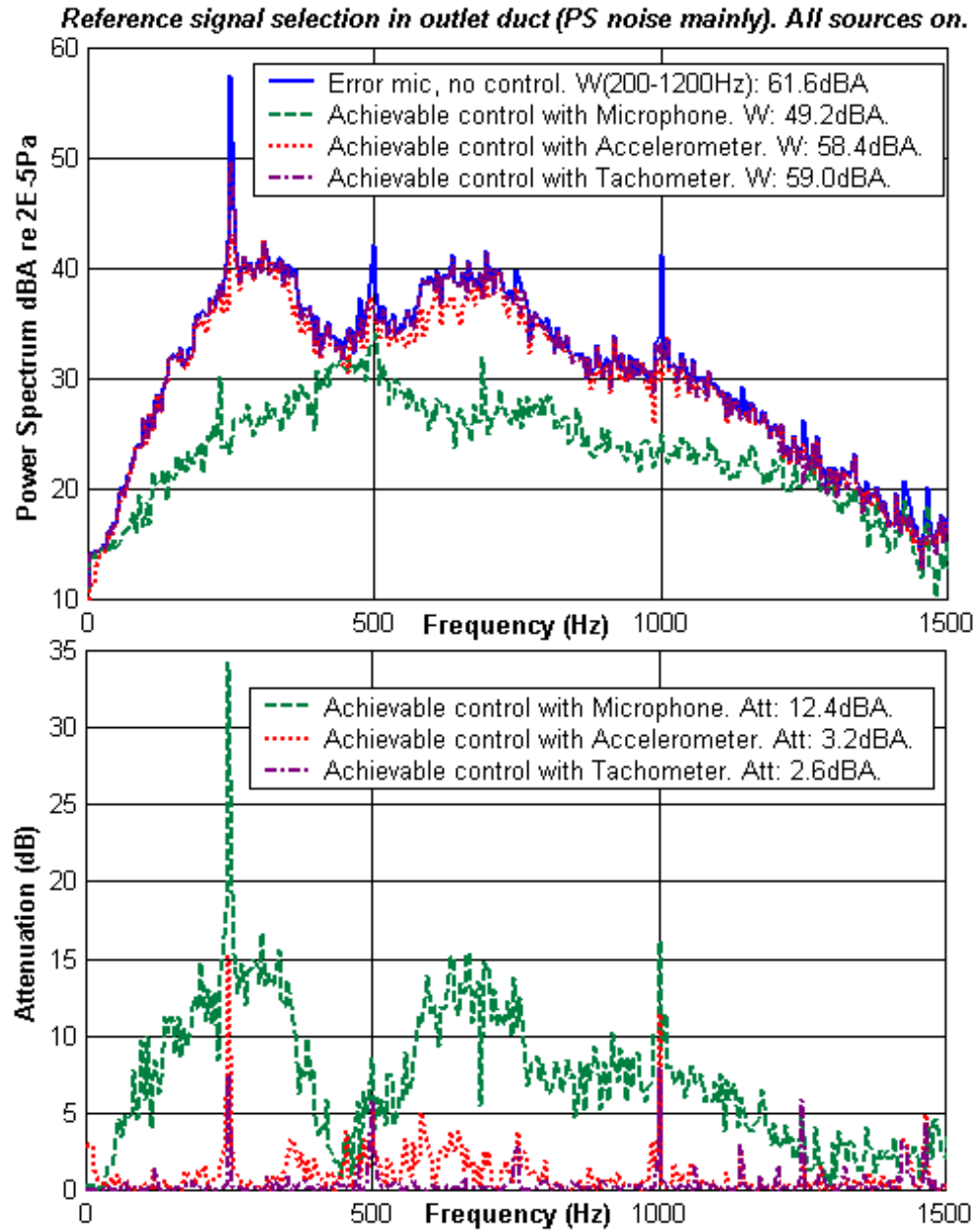


Figure 5.30 Single channel ANC in outlet duct. Reference signal selection.

5.3.2.3. Group Delay in the New Transducers.

The speaker used in the new duct prototypes is a driver of 4" diameter in a sealed enclosure. We measured the transfer function between this transducer and the microphones in the duct in order to deduce the Group Delay of the system speaker-microphone (removing the propagating lag due to the separation distance between source and receiver). The results for two measurements are shown in Figure 5.31.

We observe that, once again, the speaker presents an important Group Delay at low frequencies (below 400Hz), which could affect the controller performance. However, it will later be shown that good control is achieved over the whole frequency range. Thus, we conclude that either these Group Delay measurements do not fully represent the causality problem or that the controller can achieve a certain control even in acausal operating mode. This aspect needs further study.

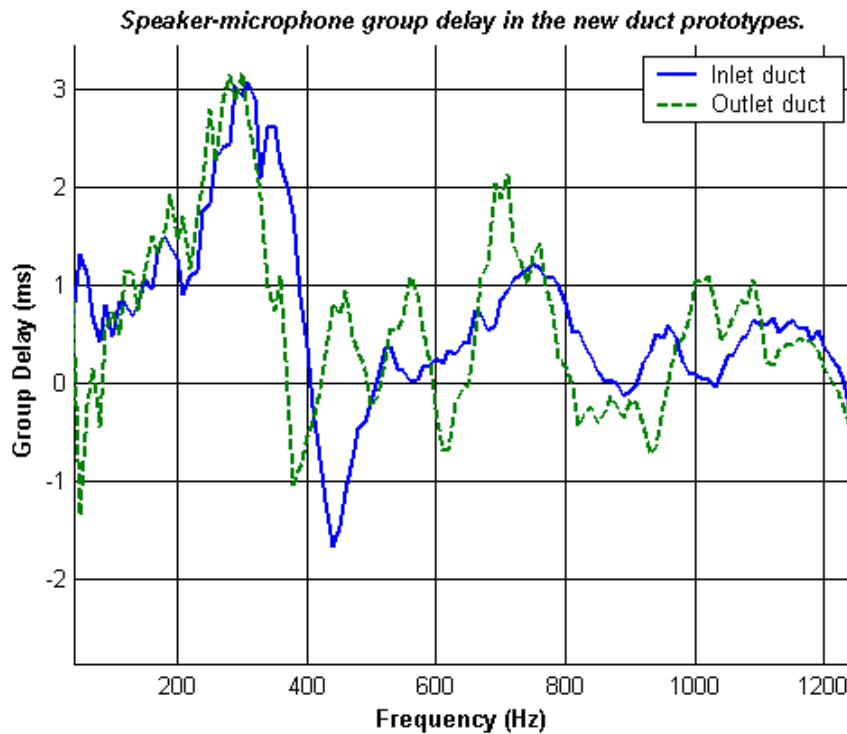


Figure 5.31 4" speaker and microphone group for the new ducts prototype.

5.3.2.4. In Duct Performance

a. Pure Tone Control

For this series of experiments, a SISO controller operating on each duct successively is used. All the noise sources present in the PC are operating but the Active Noise Control is performed in one duct at a time.

We used an accelerometer as reference signal for controlling both the inlet and outlet ducts noise radiation. As we have seen in the experiments on the first duct prototype, there is no causality issue for the control of perfectly predictable disturbance such as the fans BPF and harmonics. Thus, there is no need to minimize the delay in the control path by over-sampling the controller or minimizing the filtering of the signals. The settings for the control system are shown in Table 5.2 and the control results at the error microphones are shown in Figure 5.32 and Figure 5.33 for the inlet and outlet ducts respectively.

<i>Settings</i>	<i>Inlet duct controller</i>	<i>Outlet duct controller</i>
Reference signal analog filter:	Bandpass 100-800Hz	Bandpass 100-1250Hz
Control signal analog filter:	Lowpass 800Hz	Lowpass 1250Hz
Error signal analog filter:	Bandpass 100-800Hz	Bandpass 200-1250Hz
Controller sample rate:	2500Hz	4000Hz
Digital identification filter:	256 taps	256 taps
Digital control filter:	256 taps	256 taps
Feedback removal filter	No	No

Table 5.2 SISO controller settings for pure tone control

One important observation is that, for each duct, the noise generated by the local source (i.e. power supply block fan for the outlet duct and processor fan for the inlet duct) is well controlled. On the other hand, each source contributes not only to the noise of the duct it is connected to but also to the other duct. This “cross-term” contribution is not controlled in neither of the ducts because the unique reference sensor used does not characterize both sources but only the one it is attached to. As expected, no broadband noise is controlled and there are no causality issues.

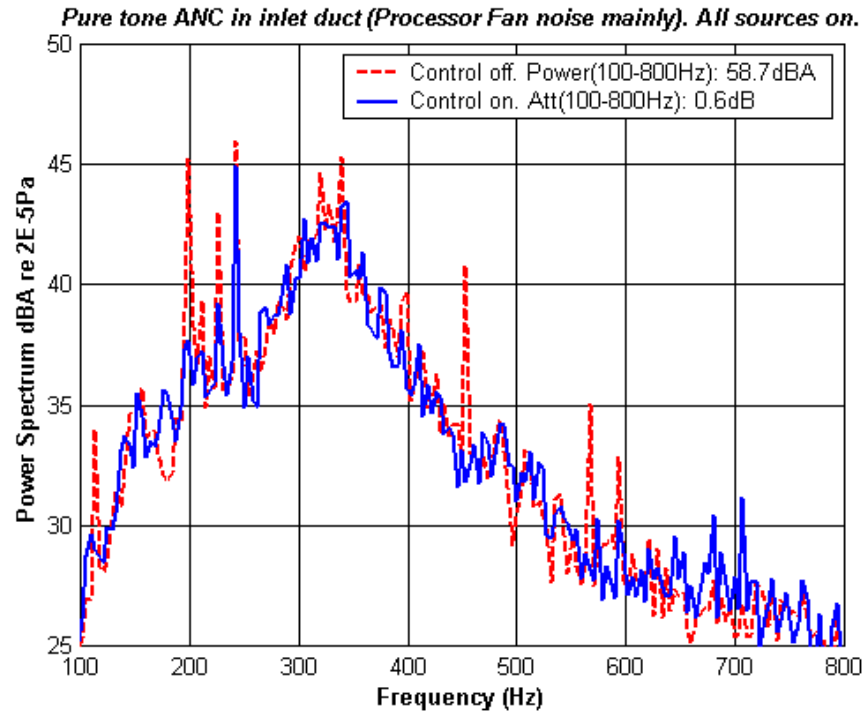


Figure 5.32 Pure tone ANC in new inlet duct prototype. Error microphone level.

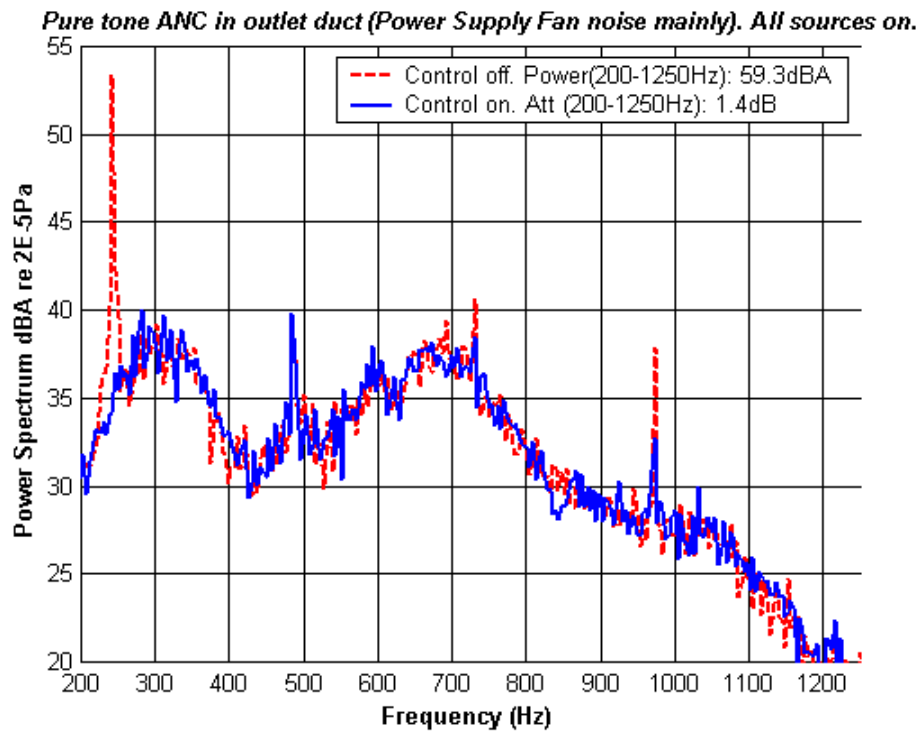


Figure 5.33 Pure tone ANC in new outlet duct prototype. Error microphone level.

b. Broadband Noise Control.

For the control of broadband noise radiation, we need to use a pressure sensor as reference signal. Since the noise is non predictable, we are facing causality problems again. Thus, the new controller settings are optimized to minimize the delay in the control path as can be seen in Table 5.3.

<i>Settings</i>	<i>Inlet duct controller</i>	<i>Outlet duct controller</i>
Reference signal analog filter:	Highpass 80Hz	Highpass 80Hz
Control signal analog filter:	Lowpass 3150Hz	Lowpass 3150Hz
Error signal analog filter:	Bandpass 80-1250Hz	Bandpass 80-1250Hz
Controller sample rate:	7500Hz	7500Hz
Digital identification filter:	80 taps	80 taps
Digital control filter:	80 taps	80 taps
Feedback removal filter	Yes	Yes

Table 5.3 SISO controller settings for broadband noise control.

Results in Figure 5.34 and Figure 5.35 show that good control at the error microphone can be achieved above 200Hz. We observe a spillover phenomenon at lower frequencies, which is characterized by an increase rather than decrease in the sound pressure level in this frequency range. The control in the outlet duct is operating almost ideally above 400Hz (causal control), which is not the case for the inlet duct. Also, many tones are not controlled in the inlet duct. We believe this important residual noise is correlated to the power supply block fan and / or hard disk drive. In the next experiments, we perform a two input / two output control on both ducts simultaneously which could help to improve the noise reduction in the inlet duct. Also, it will give us the opportunity to look at the overall performance of the global ANC system in terms of total sound power level radiated by the PC.

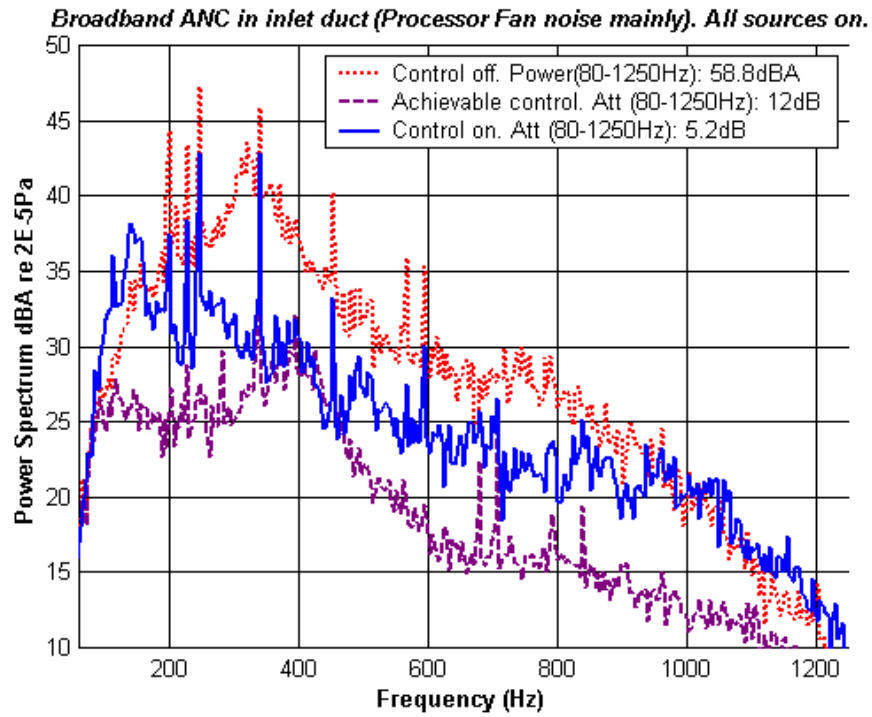


Figure 5.34 Broadband ANC in new inlet duct prototype. Error microphone level.

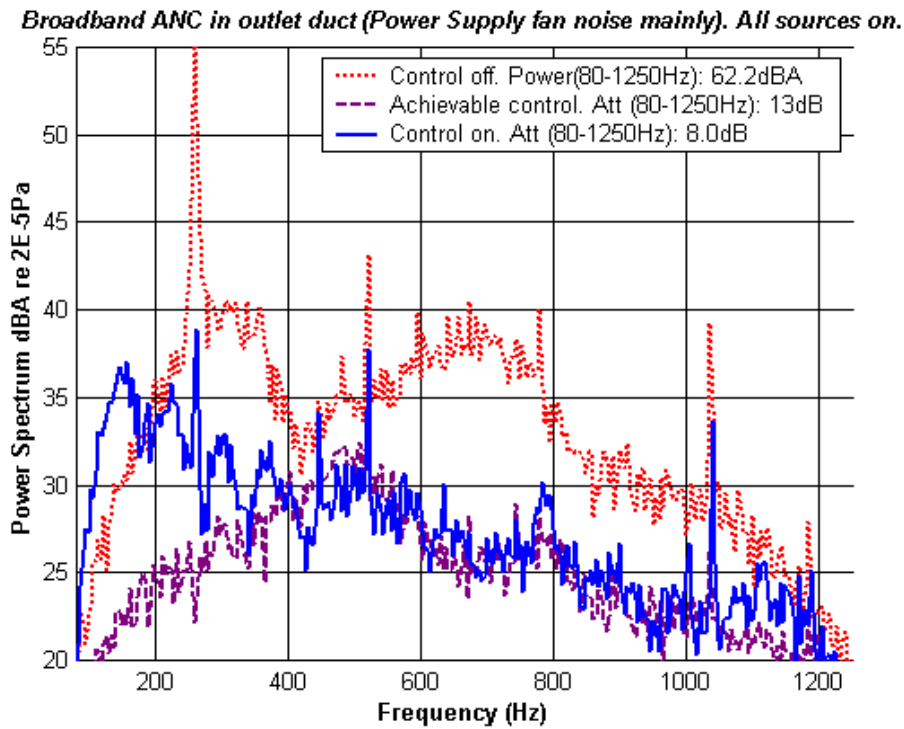


Figure 5.35 Broadband ANC in new outlet duct prototype. Error microphone level.

5.3.2.5. Overall Performance

a. Controller Setup

In order to measure the overall performance of our global active noise control system, we need to control the noise radiated by the inlet and outlet ducts simultaneously. Thus, the controller used here is slightly more complex than earlier. We use a multiple input / multiple output digital feedforward controller based on the Filtered-X LMS algorithm. The controller is implemented on a Texas Instruments C40 board, also interfaced via a Labview code.

The MIMO control uses two reference sensors (microphones located near the primary noise sources in the ducts), two control signals (for each speaker in the ducts) and two error signals (microphones at the ducts outlet). The system is considered fully coupled. It is thus assumed that both primary sources can affect the response at both error sensor locations. Although such strategy leads to an increased computational effort compared to the independent controllers case, it is a necessity when there is indeed “cross-term” coupling in the system. One other observation is that the cost function now corresponds to the sum of the instantaneous mean square errors. For more details about the multiple channel control system, one should refer to section 2.1.3.

The block diagram of the 2I2O controller designed for this series of experiments is shown in Figure 5.36. $\mathbf{P}$ is the (2 by 2) primary path transfer function matrix associating the two reference signals to the two error signals. $\mathbf{W}$ is the (2 by 2) control filter matrix composed of four adaptive filters generating two control signals using the two reference signals. $\mathbf{S}$ is a model of the secondary path, composed of four filters relating the two control signals to the two error signals.

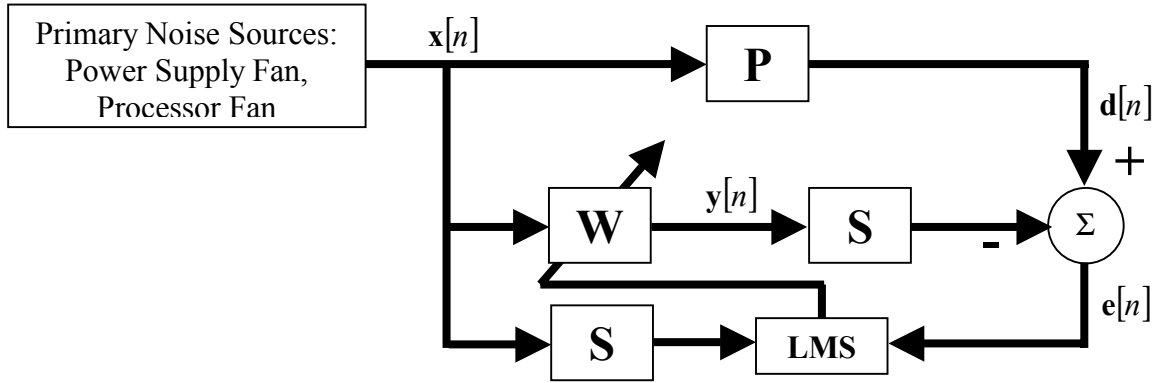


Figure 5.36 Block diagram of the 2I2O controller for global duct ANC.

As we have seen in section 2.3.2.2.b, the multiple channel Filtered-X LMS algorithm is given by the formulation below:

$$w_{jil}[n+1] = w_{jil}[n] + \mu \cdot \sum_{k=1}^K r_{kji}[n-l] \cdot e_k[n], \quad (5.14)$$

where:

- $r_{kji}[n]$ is the i^{th} reference signal filtered by s_{kj} , the estimate of the secondary path between the j^{th} control source and the k^{th} error sensor.
- $e_k[n]$ is the k^{th} error signal.
- w_{jil} is the l^{th} tap of the control filter between the i^{th} reference signal and the j^{th} control source.

We present below the results obtained with the implementation of a two input / two output multiple channel feedforward adaptive controller using the multi-channel version of the Filtered-X LMS algorithm for the control of broadband noise radiation of both ducts simultaneously. The controller settings are listed in Table 5.4.

Even though the Texas Instruments C40 board is more powerful than a C30 board, due to the larger number of control filters (four), the sample rate of the controller cannot be as high as earlier. More taps than before are needed to model the “cross-paths” from one duct to the other.

<i>Settings</i>	<i>Inlet duct</i>	<i>Outlet duct</i>
Reference signal analog filter:	Bandpass 20-3150Hz	Bandpass 20-3150Hz
Control signal analog filter:	Lowpass 3150Hz	Lowpass 3150Hz
Error signal analog filter:	Bandpass 20-1250Hz	Bandpass 20-1250Hz
Controller sample rate:	5000Hz	
Digital identification filter:	120 taps	120 taps
Digital control filter:	120 taps	120 taps
Feedback removal filter	No	No

Table 5.4 2I2O controller settings for broadband noise control.

b. ANC Results

For the following experiments, the PC, equipped with the new inlet and outlet duct prototypes, was located in a VAL anechoic chamber, placed over a reflecting board in order to perform sound power level measurements while operating the 2I2O ANC system. Thus, we could not only check the performance of the controller at the error sensors location but also look at the control effect on the global noise radiation of the PC.

As can be seen in Figure 5.37 and Figure 5.38, good noise control is achieved at the error microphones that are located at the two ducts outlet. We also show the achievable control based on the multiple coherence function between the reference signals and the error signal in each duct. For more details about partial coherence analysis, one should refer to section 2.1.3. Contrary to the SISO control, contributions of both primary noise sources are controlled in each duct. For both ducts, we observe an increase in noise level at the lower frequencies. This phenomenon is referred as spillover.

Figure 5.39 through Figure 5.41 show the effect of the Active Noise Control system on the overall sound power level radiated by the ducted PC when all the sources, only the power supply block fan, and only the processor fan are operating respectively. The results seem to be not very satisfactory. The pure tones are not attenuated and broadband noise reduction is achieved over a limited frequency range, dependent on the primary source complexity.

The best results are achieved when the processor fan is the only noise source operating (see Figure 5.41). We recall that the processor fan is, in this case, structurally decoupled from the inlet duct. Thus only airborne noise radiation paths are remaining. We observe in Figure 5.41 that broadband noise is well attenuated in some frequency bands but not others. Pure tones have their level either unchanged or increased. A similar phenomenon is observed with the power supply noise source (see Figure 5.40) or when all the primary noise sources are operating (see Figure 5.39). It is important to note that these observations concern the global sound pressure level. The in duct control is still very good for each of the three studied cases.

One suggestion regarding the limited global effect of the control system, even though the in duct control is good, is that, once the ducts are located on the PC, the major noise radiation path might not still be the air inlet and outlet. Indeed, when investigating passive noise control treatments, we found that the ducted PC casing still presented a low noise transmission loss due to both the presence of important leaks in the casing assembly as well as to the physical properties of the casing panels. The fans radiate energy outside the PC casing (actively controlled path) as well as inside the casing (not controlled). Also, our secondary noise sources radiate not only downstream the ducts but also upstream (thus into the PC casing). Thus, the total energy level inside the PC casing is either unchanged or increased when the active noise control system is operating. Since the casing presents a poor noise transmission loss, much of the overall noise radiated by the ducted PC might be coming from airborne noise transmission through the PC casing panels and leaks.

The other suggestion is that the error microphones, being located in the near field of the secondary sources, might control the near field high order modes of the secondary noise source rather than the plane wave propagating disturbance from the secondary source. At the time this thesis is written, new work is being conducted to find out what are the causes of the limited overall performance. In terms of feasibility, it was proved that it is possible to achieve good broadband noise attenuation in a fairly small sized duct using a simple SISO Active Noise Control system. In order for the Global ANC system to be effective, some noise control applied to the PC casing is likely to be necessary.

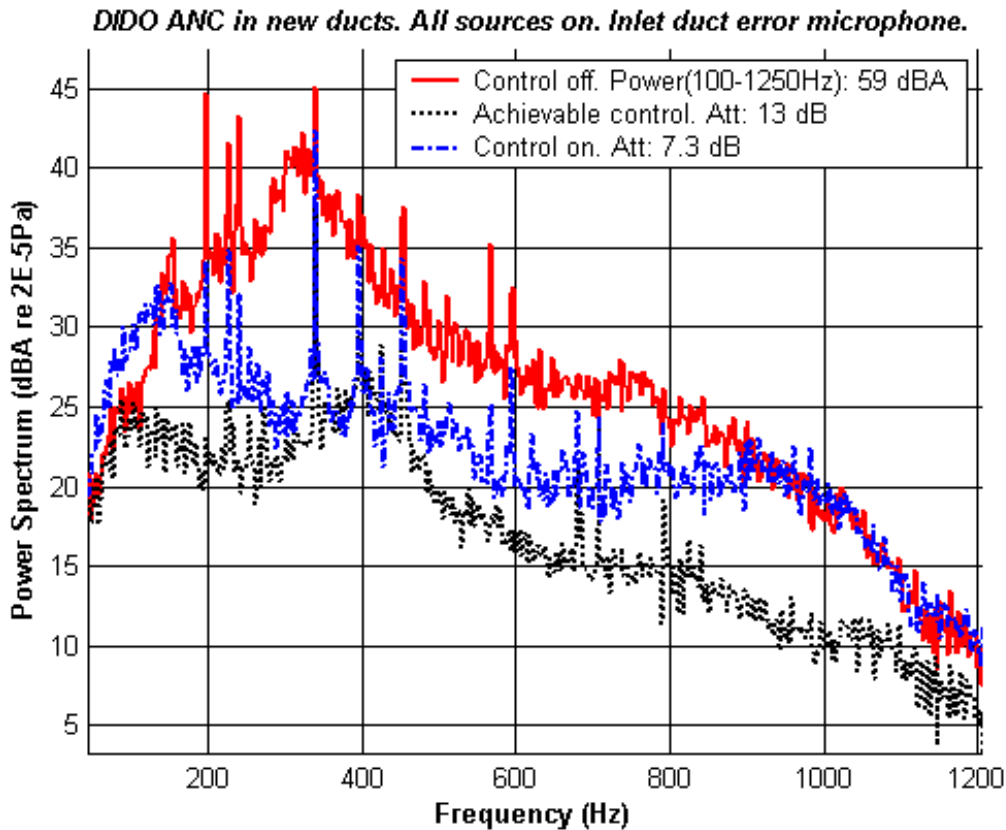


Figure 5.37 2I2O ANC in new ducts. All sources on. Inlet duct microphone level.

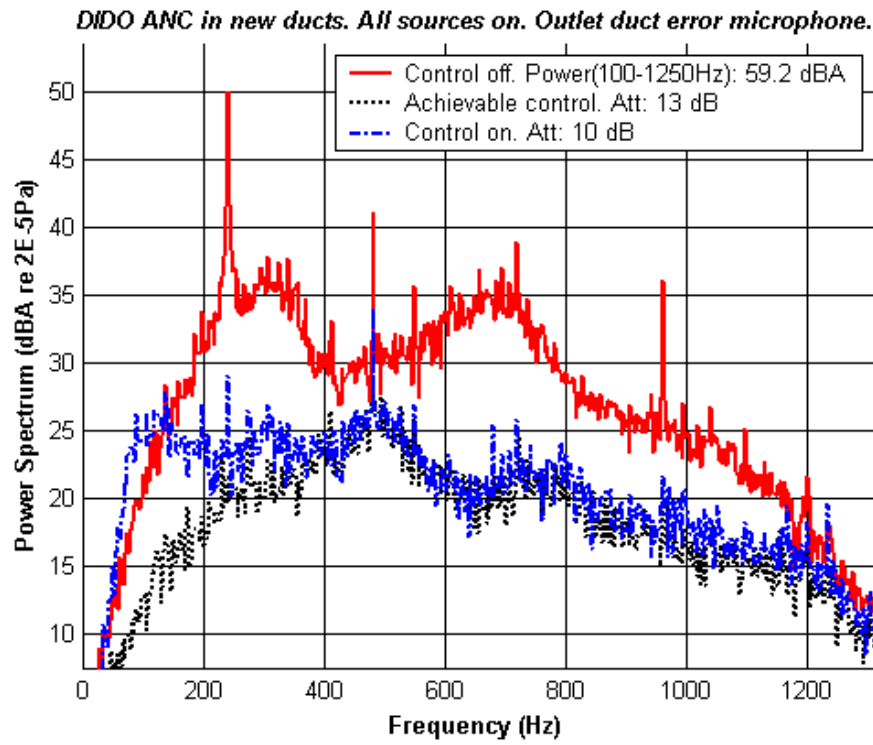


Figure 5.38 2I2O ANC in new ducts. All sources on. Outlet duct microphone level.

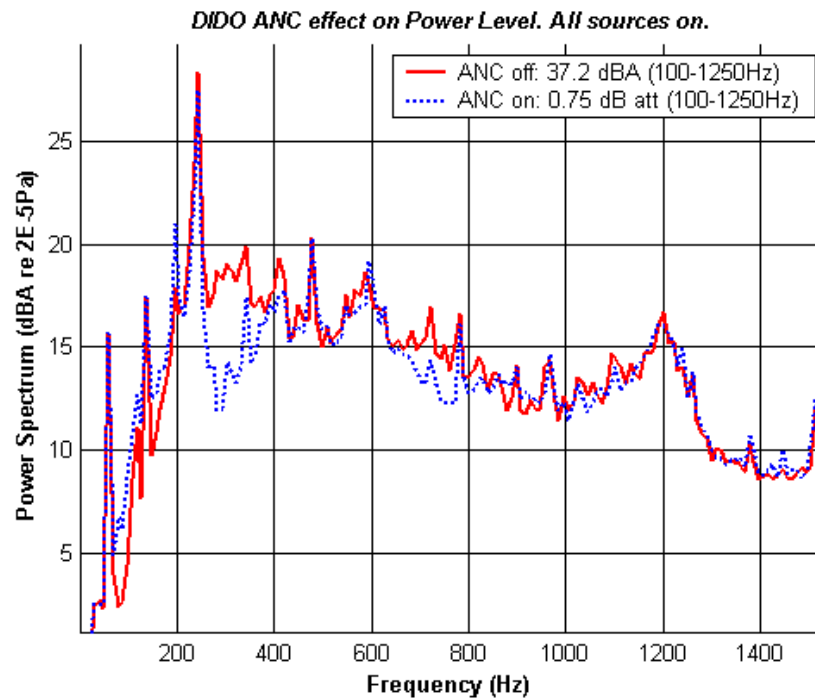


Figure 5.39 2I2O ANC effect on sound power level. All sources on.

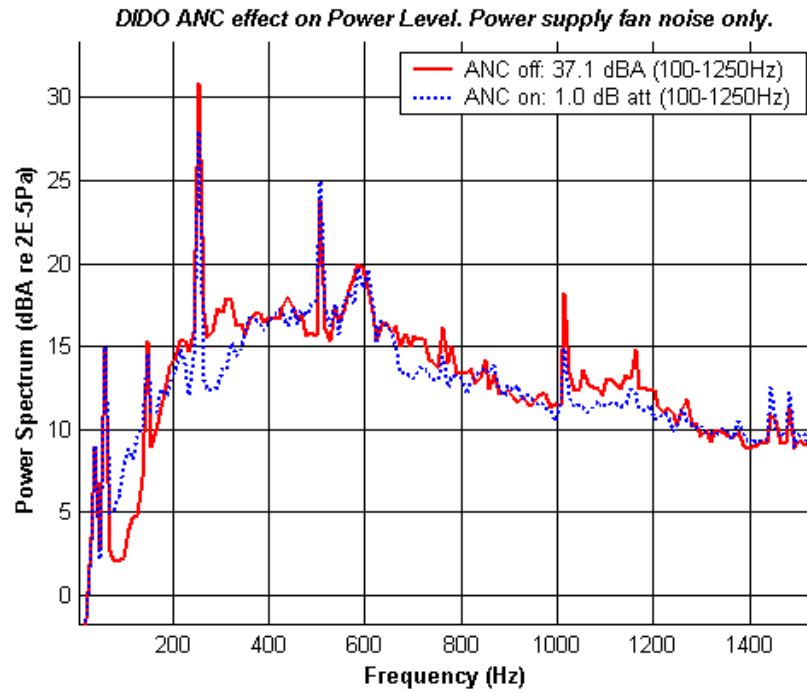


Figure 5.40 2I2O ANC effect on sound power level. Power supply fan noise.

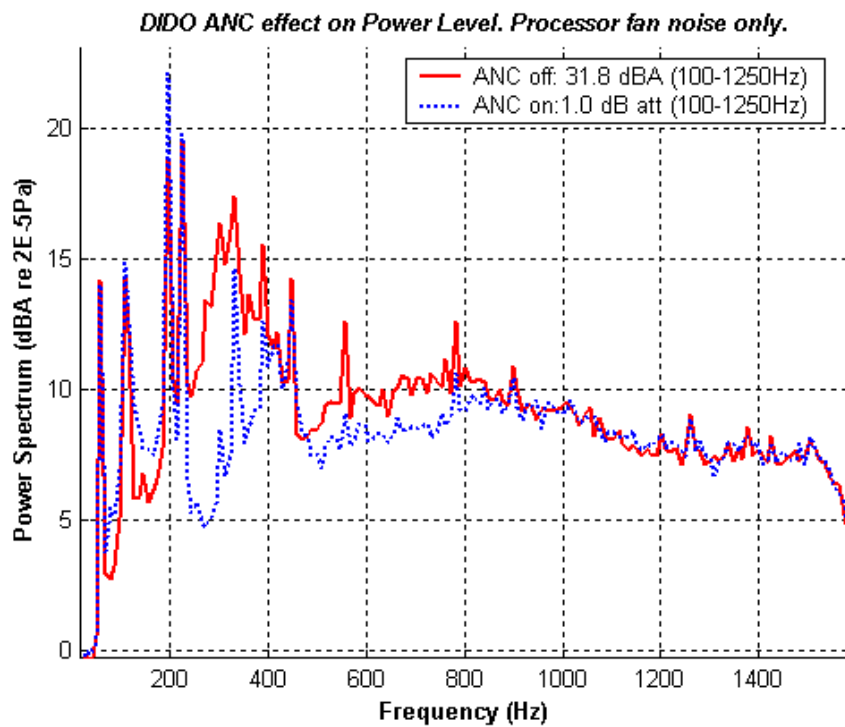


Figure 5.41 2I2O ANC effect on sound power level. Processor fan noise.

Chapter 6. Local Active Noise Control

Local Active Noise Control is the other approach we investigated for the attenuation of noise radiated from Personal Computers. As opposed to the global control strategy, we do not try here to cancel the noise at its source but rather at the receiver location, where the PC user is sitting. The major advantage of the local control system is that it does not require any redesign of the PC unit in order to be implemented. The idea is to make use of the inboard multimedia equipment of the PC in order to implement a cost effective ANC system. Similarly to the global control approach, we will limit the field of study to the active control of noise. The work on reduction of vibrations for noise control purposes has been purely passive. Thus, pressure and vibrations sensors will be used in conjunction with loudspeakers in this local ANC strategy.

The controllers used for this series of experiments are the same as those used for the global ANC. Thus, we will not explain in detail their principle of operation in this chapter. For more information about the single input / single output and multiple input / multiple output feedforward digital controllers and the filtered-x LMS algorithm, one should refer to section 2.3. The focus here will rather be on the work carried out to maximize the controller performance, and present the achieved results in terms of attenuation at the error sensor as well as size of the zone of quiet created.

6.1. Control Setup Presentation

6.1.1. Single or Multiple Channel Controller?

The work on Local ANC was performed in two consecutive phases. In a first series of experiments, a SISO control architecture is investigated, with a single secondary noise source controlling the noise level at a unique pressure sensor (microphone) location. Since the ultimate goal is to create a reasonably large zone of quiet around the user's head, it will be necessary to use multiple error sensors. Thus, the second part of the work consists in implementing a multiple channel controller.

In the global ANC system, the control work has focused on the airborne noise dominated by a single primary noise source: the power supply block fan noise in the outlet duct and processor fan noise in the inlet duct. In the present case, the goal is to control the noise radiated by the whole PC, which includes multiple radiation paths and primary noise sources. Typically, no unique sensor is a perfect reference for all the primary sources. Thus, there is usually at least one sensor per primary source operating. Also, in a local control strategy, the primary disturbance acoustic field is usually complex due to the presence of multiple non-collocated primary sources and their interaction with the environment. Thus, several control sources are usually needed to control the sound field at multiple locations in space. These reasons justify the ultimate use of a MIMO controller rather than a SISO controller in a local noise control strategy.

The SISO and MIMO controllers used for the following experiments are the same as those used in the global control strategy. Both types of feedforward digital controllers incorporate the Filtered-X LMS algorithm to update the control filters. For more information about the Texas Instruments DSP boards and Labview based hardware interface constituting the VAL Labs controllers, one should refer to section 5.2.2.

6.1.2. Implementation in a Personal Computer

A block diagram representing the typical implementation of the local active noise controller in a PC is shown in Figure 5.1. In the most complete control configuration, microphones and accelerometers located on the vibrating noise sources identified in Chapter 3 (power supply block fan, processor fan, and eventually hard drive) will be used to provide information about the primary disturbance prior-to-control. On board multimedia speakers can be used as secondary noise sources, provided they are adequately located between the PC and the user (we will have to deal with causality issues, similarly to the global ANC problem) and also present a large enough source strength.

Modern computers are usually equipped with microphones and speakers as well as sound cards that integrate D/A and A/D converters and often DSP chips. Our work is limited to a study of feasibility of local ANC for personal computers. We will be using external digital controllers implemented on DSP hardware that allows real time sound analysis and synthesis. Ultimately, the goal is to integrate the noise control hardware in the PC. This can be done by integrating the DSP boards into the machine or by using DSP chips eventually present in the standard sound card. The other idea would be to use the computer's main board computational resources by performing the control using the CPU, as a background task. Those issues, beyond the scope of this thesis, are currently under investigation.

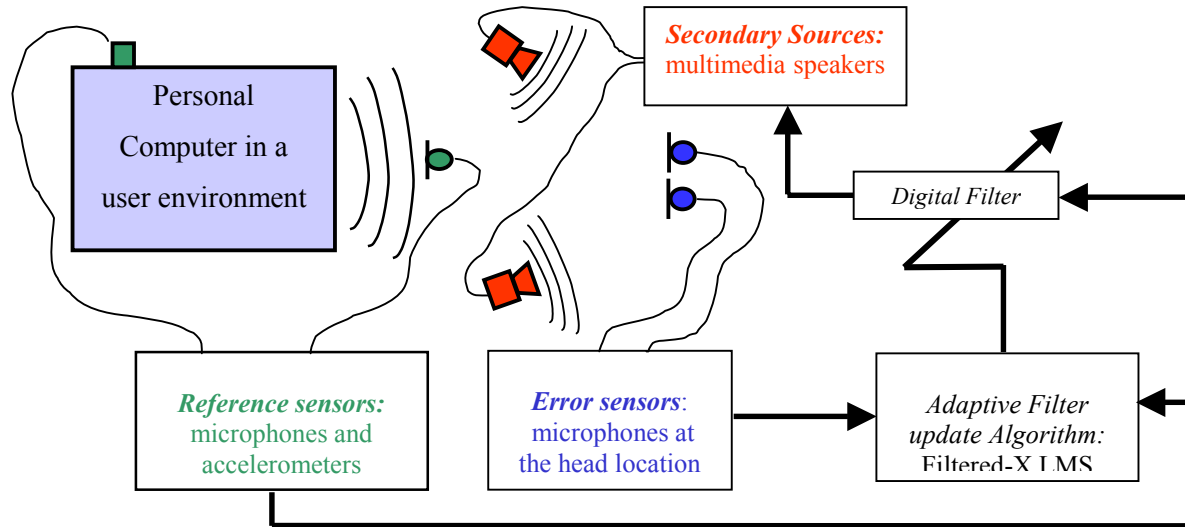


Figure 6.1 Block diagram of a local ANC system for controlling noise of PC's.

6.1.3. Operating Environment.

We expect the computer to be located in a bedroom, office, or living room, all constituting semi-reverberant environments. For our experiments, the computer is located in one of the VAL anechoic chambers. A semi-reverberant environment is simulated by placing several reflecting surfaces representing for example the floor, side, and rear walls. There are three advantages to such setup:

1. The background noise is kept very low since any sound is absorbed when reaching an anechoic wall. This is important to obtain the best control since no extraneous noise at the error sensors can limit the system performance.
2. The proportion of direct and reflected sound waves at the error sensor is controlled, which represents the complexity of the primary sound field. In turns, this determines the complexity of the secondary sound field used to cancel the primary field.
3. The setup environment is not truly an enclosure since it does not support modes. This last point is critical since the modal characteristics of an enclosed sound field are the primary parameters affecting the extension of the zone of quiet. Indeed, in an enclosed sound field, the number of control sources required to globally

control the sound pressure in theory equals the number of excited modes [9]. By placing the computer in an anechoic environment, we limit the number of “degrees of freedom” of the plant, not having to deal with the room response. In consequence, the extent of the zone of quiet will be purely determined by how well the primary and secondary direct sound fields couple.

6.2. SISO Feedforward Control Results

6.2.1. Reference Signal Selection

Our initial work on local control consisted of the implementation of a single input / single output controller. The system setup was similar to Figure 6.1 but a unique loudspeaker was used to control the noise at single error microphone. Moreover, although multiple reference sensors are shown in Figure 6.1, a unique reference signal was used in the SISO control test. Eventually, several reference signals could be mixed in order to generate a unique reference signal coherent with multiple primary noise source. However, if the propagation paths of the disturbances sensed by each reference sensor vary significantly the one from the other (particularly in terms of delay), the unique control filter is not likely to lead to the best control performance [9]. This mixing approach was not investigated for the following experiments.

6.2.1.1. Problem Simplification

For the SISO control tests, we limit the problem to a unique operating primary noise source. One of the major reasons for this simplification is that no one reference sensor can usually generate an optimal signal for multiple primary sources operating simultaneously and generating uncorrelated noise. The only case where such approach could work is if two or more sources are collocated, of similar source strength, and that the reference sensor used is not specific to one the sources (for example not a tachometer). In the case of our PC, none of the primary sources are close enough in order for us to use one sensor for two sources, although a microphone located on the rear side of the PC could eventually sense the computer fans noise. Regarding the hard drive, we

have seen in section 3.2.2.3 that it does not constitute a dominating noise source below 1000Hz. Thus, the local ANC is limited to the control of fan noise and no reference sensor is devoted to the pick up of hard drive noise and or its vibration.

For the following experiments, the PC is located in one of the VAL anechoic chambers in a semi-reverberant environment simulating a simplified room as can be seen in Figure 6.2 and Figure 6.3. Three reflecting surfaces are present as well as a desk. The absence of back wall behind the error microphone and ceiling limit the proportion of reflected sound field reaching the error microphone. In a real life environment, the noise field would most likely be less direct at the error sensor location but never dominated by the reflected field like in a reverberant room. Moreover, the user will be typically sitting next to one or two walls, with the rear wall and at least one side wall further away than the front wall. Thus, it is likely that the most important incident waves reaching the user head are the direct field and first reflections from nearby walls. Our simulated environment can thus be considered reasonably realistic.

For the control of fan noise, three types of reference sensors can be used: pressure sensor (microphone), vibration sensor (accelerometer), or tachometer. One advantage of the tachometer sensor is that is usually built in the fan units (it is the case for both the processor and power supply block fans we are working on), which reduces the cost of implementation of the controller. One other important advantage is that such reference sensor is not sensitive to the secondary sources operation, thus preventing any control instability due to feedback.

As we have seen for the global ANC strategy, each sensor provides limited information about the primary noise source, which is also investigated below. Figure 6.4 presents the achievable attenuation at the user head location when the power supply block fan is the only primary noise source operating. Figure 6.5 corresponds to the processor fan noise, which was not structurally decoupled from the PC casing for any of the local ANC experiments. Thus we observe in Figure 6.5 the presence of multiple tones characteristic of the structure-borne radiation path for the processor fan noise. Four different sensors, which are a microphone located in the near field of the fan, an accelerometer on the fan (or power supply block), an accelerometer on the PC casing and a signal from the fan's built in tachometer, are considered as reference signal.

The bandwidth of ANC is limited to the 80-1000Hz frequency range for two reasons. First, it was observed in the Chapter 3 that the fan noise is mainly present below 1000Hz. Also, the higher the frequency, the smaller the size of the zone of quiet [9]. The two reasons are that, when frequency increase, the source complexity usually increases and the acoustic wavelength decreases.

The source radiation pattern becomes more complex because, the higher the frequency, the more it is disturbed by the complex environment geometry (diffraction, and reflection phenomena [26],[27]). Since the secondary sound field is required to perfectly map into the primary sound field in order to get generalized attenuation of the sound pressure in the room, the number of actuators needed will increase with frequency.

The second problem, quite similar to the first one, is due to the fact that the acoustic wavelength decreases with frequency. Indeed, only when the primary and secondary noise sources are perfectly identical (but out of phase) and collocated that the control in the acoustic space is global [9]. In practice, the primary and secondary noise sources cannot be collocated. Thus, the zone of quiet (regions of the room where the primary and secondary sound fields perfectly map) is inversely proportional to the acoustic wavelength. At low frequencies, the primary and secondary source spacing is small compared to the wavelength and they can thus be considered collocated. The zone of quiet is then likely to be large. As the frequency increases, the sources spacing become significant, and we then observe zones of noise cancellation as well as zone of noise

amplification. The most explicit example of this phenomenon is the one of the acoustic dipole. At low frequencies, such dipole has a very low radiation efficiency because the two monopoles can be assumed collocated. However, as the frequency increases, the sound field becomes more complex with noise cancellation mostly in the mid plane [26].

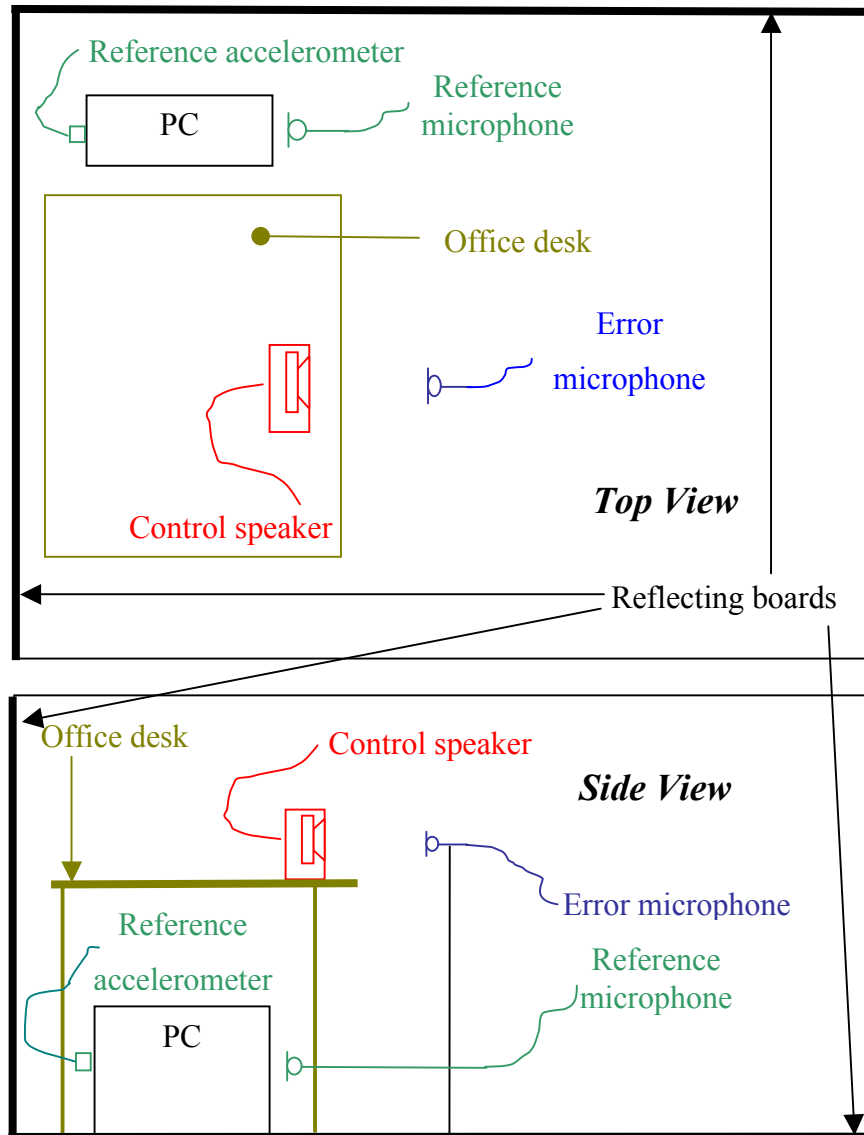


Figure 6.2 Presentation of the PC simulated user environment in anechoic chamber.

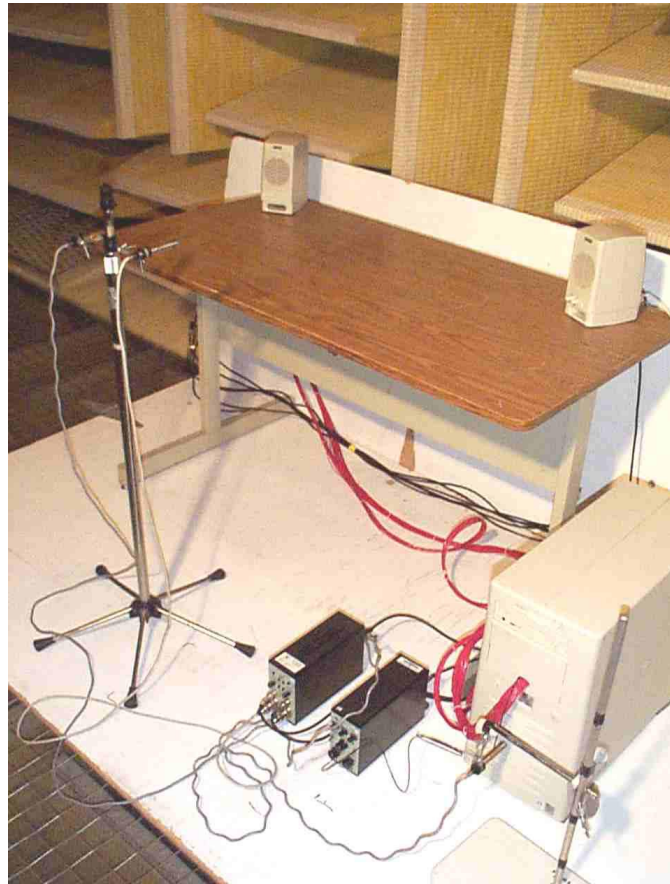


Figure 6.3 Picture the PC simulated user environment in anechoic chamber.

6.2.1.2. Reference Signal Selection for the Power Supply Block

The most theoretically achievable control performance at the error microphone, using a unique reference sensor, and when only the power supply block noise source is operating, is shown in Figure 6.4. The configuration cases were the following:

1. Top left plot: a microphone is located on the rear side of the PC casing, next to the power supply block fan outlet
2. Top right plot: an accelerometer is mounted on the power supply block casing
3. Bottom left plot: an accelerometer is mounted on the rear panel of the PC casing, next to the power supply block
4. Bottom right plot: the tachometer built in the power supply block fan is used as reference

From the graphs in Figure 6.4, we observe that any of the selected reference sensors can lead to the control of the pure tones (fan BPF at 250Hz plus its harmonics 2 and 4). On the other hand, only the sensors located in the near field of the vibrating source, i.e. the microphone and the accelerometer mounted on the power supply block, can lead to control of broadband noise. Overall, the near-field microphone is the reference sensor that can potentially lead to the largest performance with control of not only the predictable noise (fan BPF plus its harmonics) but also the broadband noise over a wide frequency range. In terms of sound power reduction, both near field accelerometer and near field microphone lead to about 9dB of overall attenuation since the error power spectrum is dominated by the pure tones contribution.

To conclude this section, the microphone is the selected reference sensor for the power supply block noise source.

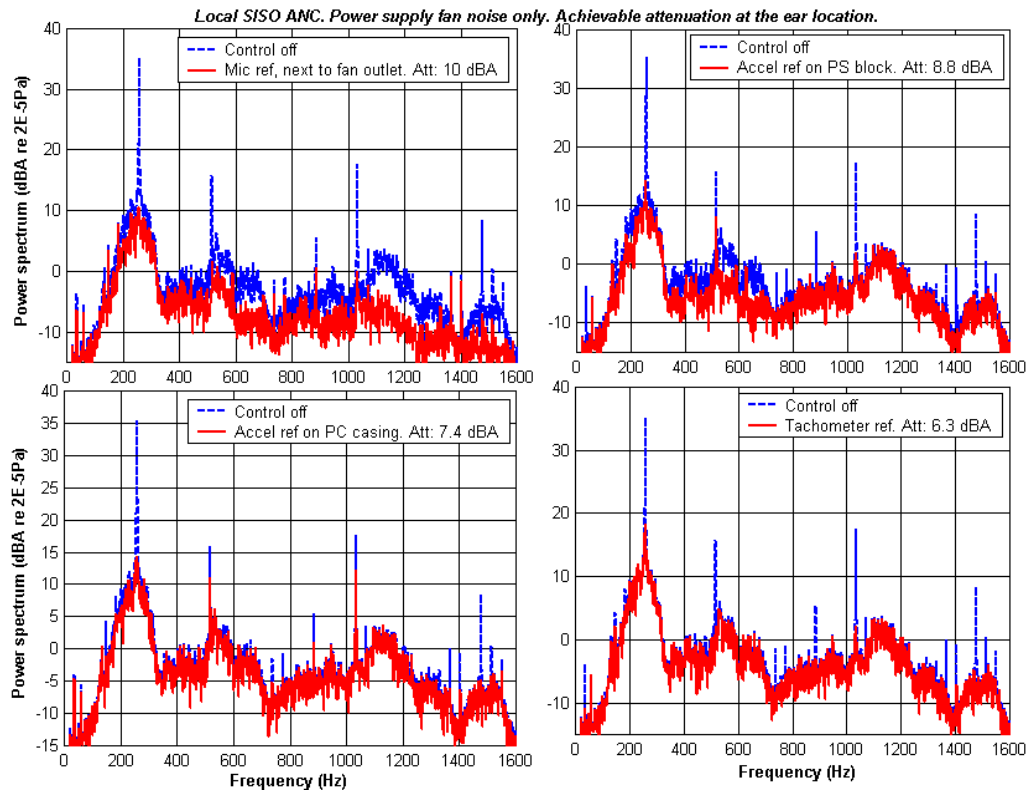


Figure 6.4 SISO local ANC of power supply fan. Reference signal selection.

6.2.1.3. Reference Signal Selection for the Processor Fan Source

The most theoretically achievable control performance at the error microphone, using a unique reference sensor, and when only the processor fan noise source is operating, is shown in Figure 6.5. The configuration cases are the following:

5. Top left plot: a microphone is located on the rear side of the PC casing, next to the processor fan outlet
6. Top right plot: an accelerometer is mounted on the processor fan housing
7. Bottom left plot: an accelerometer is mounted on the rear panel of the PC casing, next to the processor fan
8. Bottom right plot: the tachometer built in the processor fan is used as reference

From the results shown in Figure 6.5, we observe that, when the processor fan is the only operating noise source, the tachometer and the accelerometers are the only sensors that lead to the control of all the tones originating from the fan noise and/or vibration. On the other hand, the microphone located in the near field of the fan helps to attenuate only part of the pure tones but it can provide some control over the broadband radiated noise, which is not the case with the other references.

This confirms the presence of both an airborne and a structure-borne radiation path for the fan noise. While the microphone is sensitive to the airborne path, the accelerometer can mainly pick up the structure-borne portion of the radiated noise. Surprisingly, the tachometer is also an efficient structure-borne path sensor. This is probably due to the fact that the tachometer signal and the structure-borne path have both a spectrum rich in harmonics of the BPF.

To conclude this section, depending on which path (structure-borne or airborne) is to be controlled, a pressure (microphone) or vibration (accelerometer) sensor will have to be used in the following set of experiments.

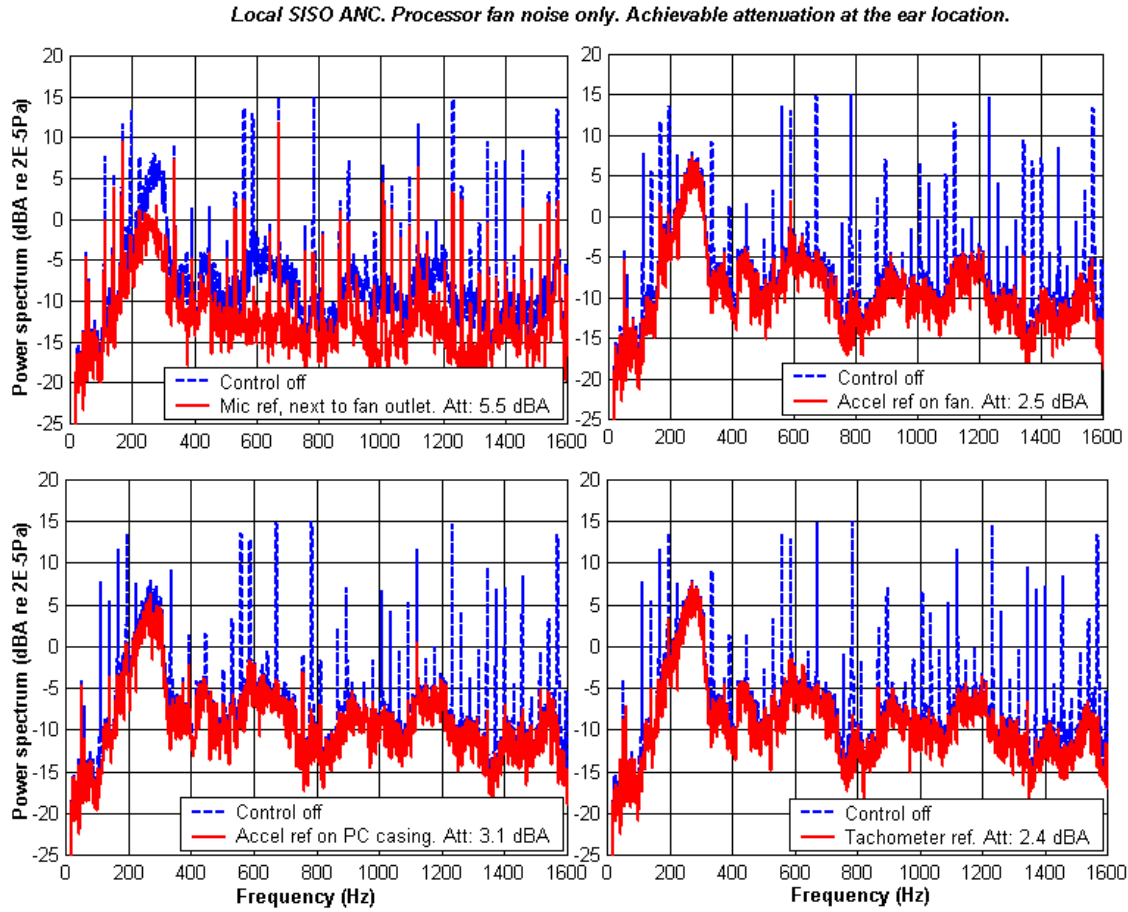


Figure 6.5 SISO local ANC of processor fan. Reference signal selection.

6.2.2. Control Results

At the time we ran the following experiments, the causality was not considered a potential source of problems. Thus, the SISO control setups shown below were not optimized to minimize the delay in the control path. The controller sample rate was set to approximately three times the highest frequency to be controlled. The frequency range of control was set to 80-1000Hz, leading to a controller sample rate of 3000Hz. The reference, control, and error signals were all bandpass filtered between 80 and 1000Hz.

6.2.2.1. Filter Size Tests

The first control parameter investigated was the selection of the number of coefficients in the control (and identification) filters. Since the controller (Texas Instruments C30 based) was sampled at a fairly low rate, a large number of taps in the control filter could be used if necessary, up to 511 taps.

a. Power Supply Block Fan Noise Results

We first consider the case of the power supply block fan noise, with a nearfield microphone used as reference. Results are shown in Figure 6.6. The graphs show the comparison between sound pressure level at the error microphone when the controller is operating (solid red line) or not (dashed blue line). The four graphs on the figure correspond to four different control tests with different control filter sizes:

1. Top left graph: 64 taps in the control and system identification filters
2. Top right graph: 128 taps
3. Bottom left graph: 256 taps
4. Bottom right graph: 512 taps

From the results in Figure 6.6, we conclude that, regardless of the number of coefficients, the same attenuation of both broadband noise and pure tones is achieved with an overall attenuation of 5 dBA, between 80 and 1000Hz, at the error microphone location. Thus, no more than 60 taps are needed to model the plant when the power supply block is the only operating source and the controller sample rate is 3000Hz.

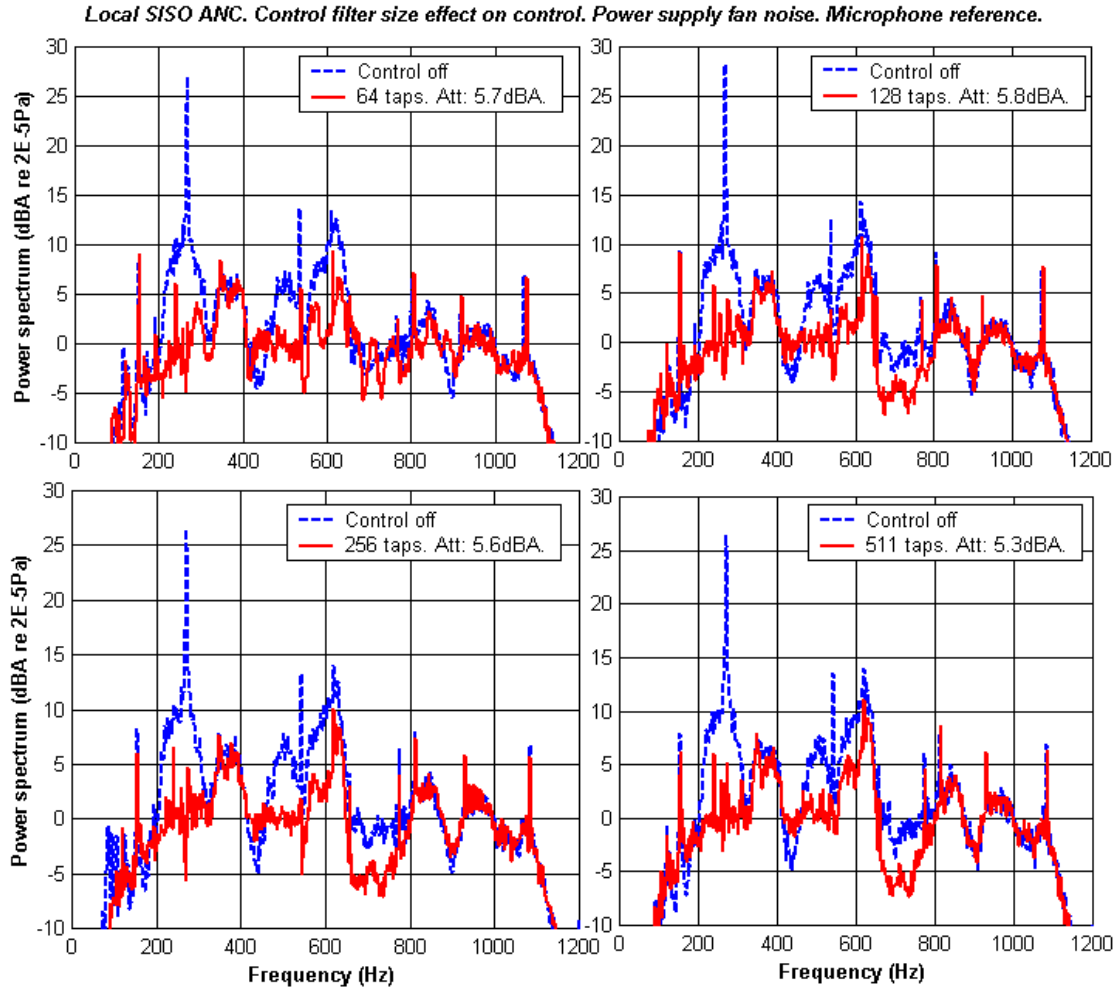


Figure 6.6 SISO local ANC of power supply fan. Effect of control filter size.

b. Processor Fan Noise Results

We now consider the case of the processor fan noise, with a nearfield accelerometer (mounted on the fan housing) used as reference. Results are shown in Figure 6.7. The graphs show the comparison between sound pressure level at the error microphone when the controller is operating (solid red line) or not (dashed blue line).

The four graphs on the figure correspond to four different control tests with different control filter sizes:

1. Top left graph: 64 taps in the control and system identification filters
2. Top right graph: 128 taps
3. Bottom left graph: 256 taps
4. Bottom right graph: 512 taps

From the results in Figure 6.7, it seems that, the larger the number of poles in the digital control filter, the higher the attenuation at the error sensor location. This observation relates to the complexity of the target control filter. In the case of “smooth” broadband noise accompanied with few pure tones such as for the power supply block fan noise, few coefficients are needed in the control filter to model the system response. On the other hand, the non-decoupled processor fan generates a spectrum rich in pure tones. Typically, one pair of filter coefficient is needed to model each pure tone (phase and amplitude information). The more tones present in the response, the more coefficients are needed in the control filter, which is why even the filter size requirements for the processor fan noise are much larger than for the power supply block fan noise.

In terms of overall noise reduction (integrated between 80 and 1000Hz), the increase in filter size does not seem as important in the case of the processor fan noise since there is only 1dB difference between the 64 taps filter case and the 512 taps situation. A summary of the overall attenuation results for the different experiments carried out is available in Table 6.1. Those attenuations correspond to integrated results, between 80 and 1000Hz, at the error microphone location. In terms of overall sound power reduction, the use of large control filter size, which leads to the use of large computation resources, does not seem to be justified in most cases. However, one has to consider that these integrated results do not take into account the factor of sound quality. For example, although many of the high frequency tones do not contribute much to the overall sound pressure level, they might be very clearly heard by user and judged particularly annoying.

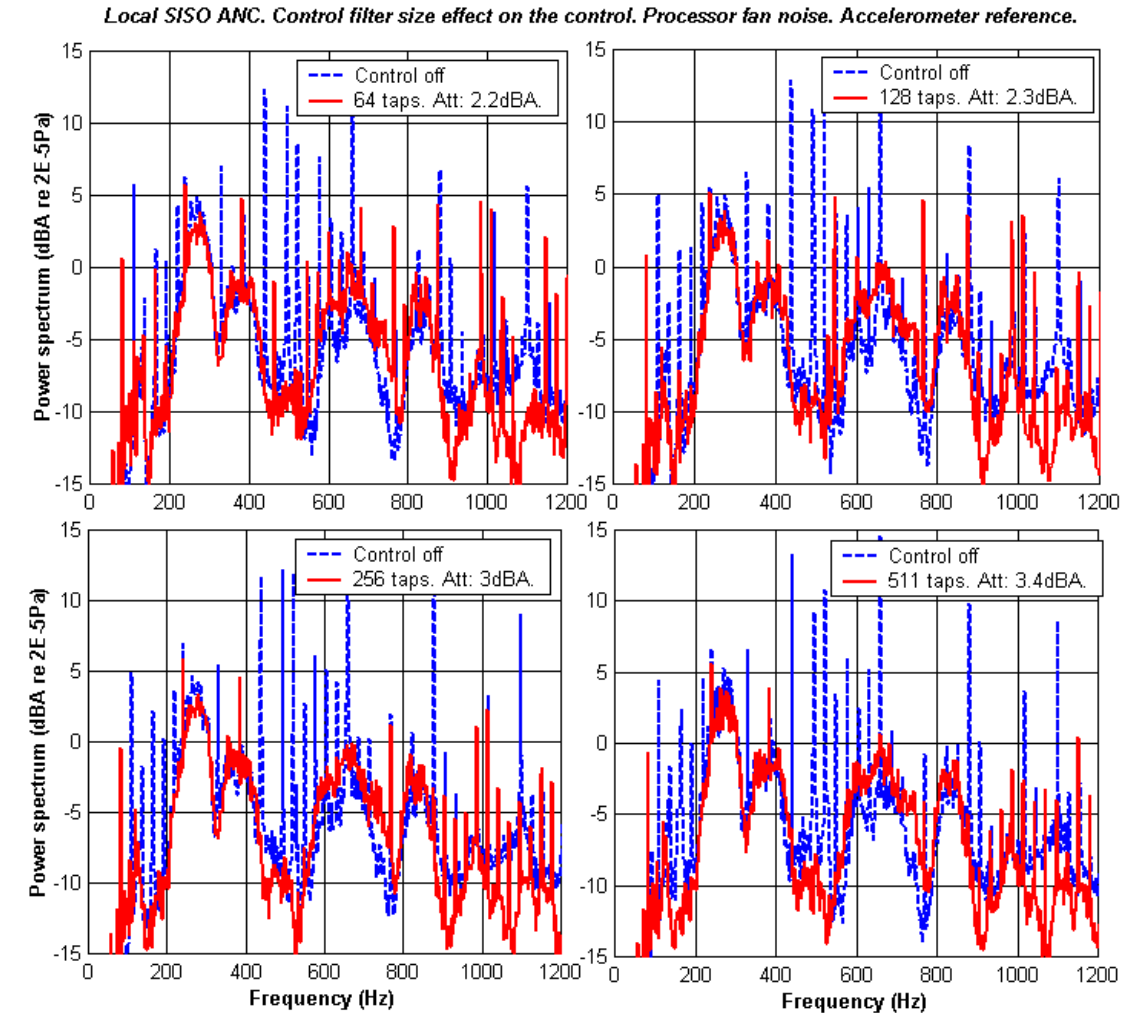


Figure 6.7 SISO local ANC of processor fan. Effect of control filter size.

Attenuation at the error microphone in dBA (in the 80-1000Hz frequency range)				
<i>Number of coefficients:</i>	<i>64</i>	<i>128</i>	<i>256</i>	<i>512</i>
Power supply fan/ Microphone as reference:	5.7	5.8	5.6	5.3
Power supply fan/ Accelerometer as reference:	3.8	4.3	4.3	4.8
Processor fan/ Microphone as reference:	3.5	4.0	4.6	4.8
Processor fan/ Accelerometer as reference:	2.2	2.3	3.0	3.4

Table 6.1 SISO local ANC performance. Effect of control filter size.

6.2.2.2. Feedback Removal and Signal Differentiation Tests

a. Test setup

Some other control parameters were investigated for the SISO Local ANC. One is the possibility to add a feedback removal filter in the control system. As explained in section 2.3.2.1.e, feedback effects can occur when the reference sensor is sensitive to the secondary source operation [9]. This can happen when a microphone, used as reference sensor, picks up signal generated from the control noise source. Such a feedback loop can lead to an instable control. In order to prevent it, the control filter can be designed in order to compensate for the feedback effect by including an inverse model of the feedback path. The feedback path is modeled at the same time as the secondary path, that is, prior to control. During this identification phase, the secondary noise source is fed with a white noise while the signals at both the error and reference microphones are recorded. A LMS-based algorithm models the secondary path “control to error” and feedback path “control to reference” by minimizing the difference between a synthesized signal and the actual measured signals. The LMS algorithm updates the coefficients of the digital filters, which operate on the control signal in order to synthesize the virtual error and reference signals. For more details about these tools, one should refer to section 2.3.

The other control option available in the VAL controller system is the differentiation of the error signal, which leads to a control that focuses on the higher frequencies (since the frequency weighting is $j \cdot \omega$ with ω the frequency in radians).

For both these tests, the computer was set-up in the regular configuration. A microphone was used as reference signal. The control and identification filters were both 256 taps FIR filters with a controller sample rate of 3000Hz. The bandwidth of control was limited to the 80-1000Hz frequency range and only one primary noise source was operating, either the power supply block fan or the processor fan.

b. Control Results

None of the options described earlier lead to a better control of processor fan or power supply block fan noise as shown in Table 6.2. The improvements observed for the power supply block fan noise could not be repeated. We believe that the feedback path removal filter is not necessary in such a control configuration due to the setup itself. Indeed, the reference microphone is located in the near field of the primary noise source and far field of the secondary noise source (at least 1 meter away). The setup does not permit strong reflections of the secondary signal toward the primary noise source. Thus, it is very unlikely that, at the reference microphone location (rear side of the PC, next to the fans), the control signal is of significant level compared to the primary noise source signal.

<i>Attenuation at the error microphone in dBA (in the 80-1000Hz frequency range)</i>				
<i>Control parameter:</i>	<i>Feedback filter off</i>	<i>Feedback filter on</i>	<i>Original error signal</i>	<i>Differentiated error signal</i>
<i>Power supply fan / Microphone as reference:</i>	3.6	4.6	4.1	6.2
<i>Processor fan/ Microphone as reference:</i>	4.3	4.5	4.1	4.6

Table 6.2 SISO local ANC performance. Effect of extra parameters.

6.2.2.3. Secondary Source Location Tests

a. Test setup

The last set of experiments with the SISO system was performed to check the influence of the secondary noise source location on the control performance. The control speaker was placed on the desk at three different locations as shown in Figure 6.8. A microphone was used as reference signal. The control and identification filters were both 256 taps FIR filters with a controller sample rate of 3000Hz. The bandwidth of control was limited to the 80-1000Hz frequency range and only one primary noise source was operating, either the power supply block fan or the processor fan.

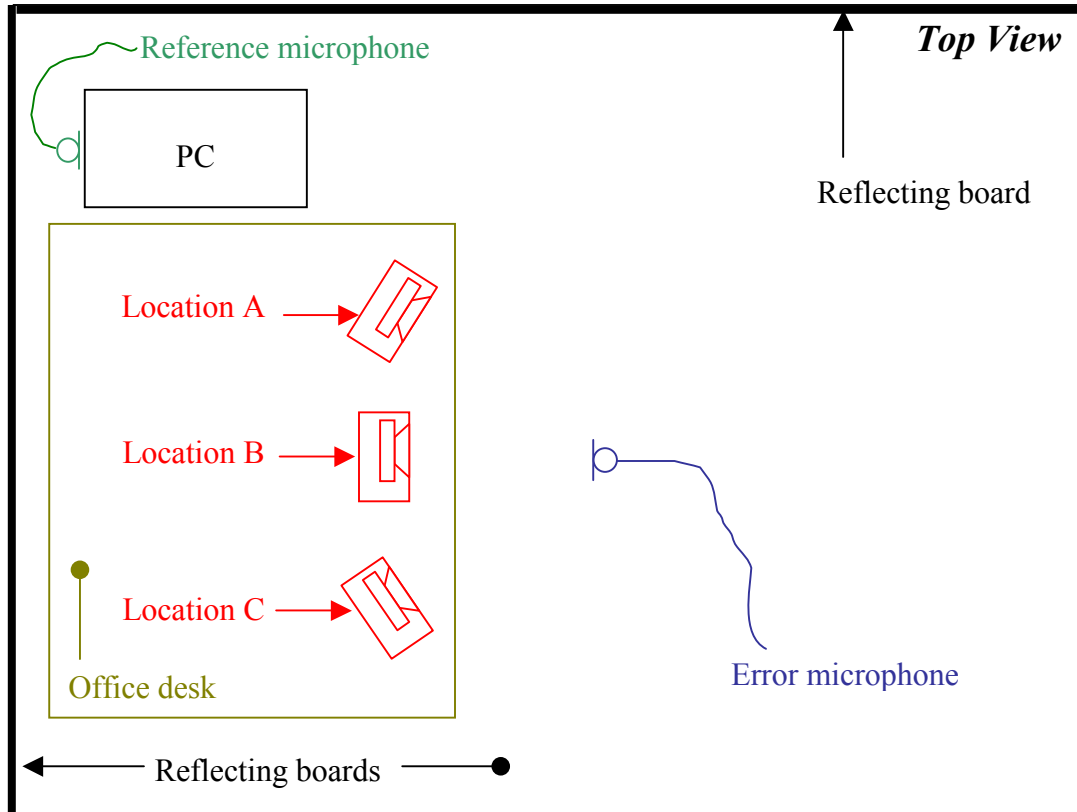


Figure 6.8 Speaker location test setup.

b. Control Results

The control results obtained at the error microphone location are shown in Table 6.3 and the attenuation spectrum obtained for the control of power supply block fan noise is presented in Figure 6.9. From the results in Table 6.3, it seems that the speaker location has an effect on the control performance only when the disturbance is broadband, which is the case of the power supply block fan noise. When the disturbance is predictable, which is the case of the processor fan noise, dominated by pure tones, the secondary source location does not affect the control performance.

From the results in Figure 6.9, it seems that the best broadband noise control results are achieved when the secondary source is the closest to the error microphone. These observations suggest that the secondary source location affects the causal performance of the system. In the case the predictable sound field, the secondary source location is not important. But, for broadband noise, it becomes important that the system be causal. In the present case, location B (see Figure 6.8) is the one for which the delay in the primary path (from the reference sensor to the error sensor) is much smaller than the delay in the control path (from the secondary source to the error sensor).

Attenuation at the error microphone in dBA (in the 80-1000Hz frequency range)			
Control speaker location:	A	B	C
Power supply fan / Microphone as reference:	6.3	7.2	6.0
Processor fan/ Microphone as reference:	4.3	4.2	4.4

Table 6.3 SISO local ANC performance. Effect of speaker location.

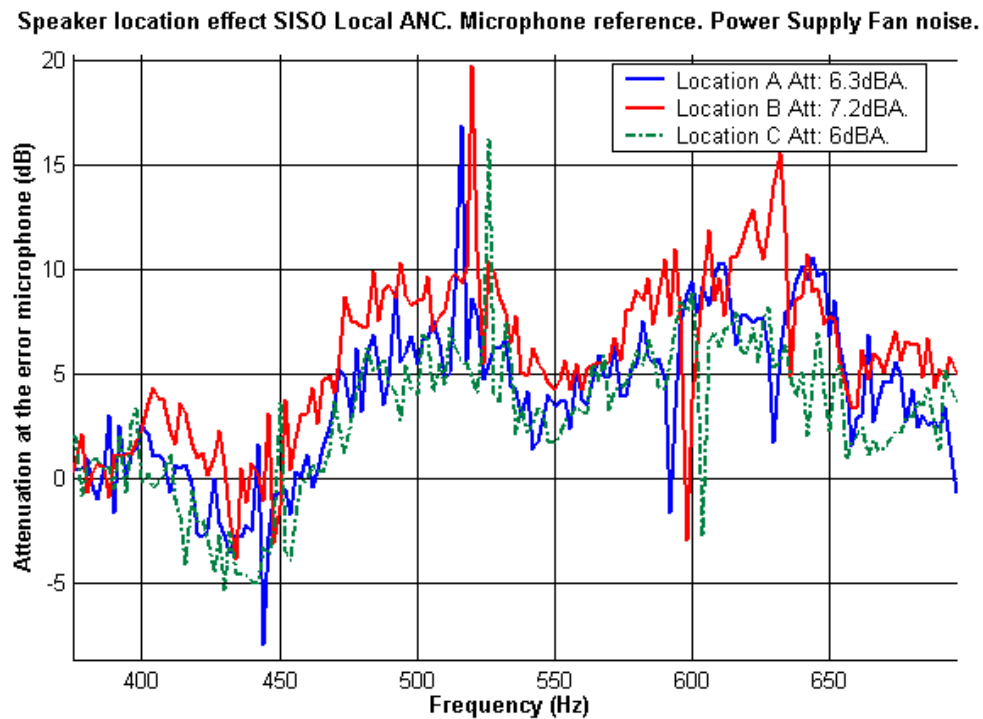


Figure 6.9 SISO local ANC of power supply fan. Effect of speaker location.

6.2.2.4. Best Performance

a. Setup

Based on the control parameters that lead to the best previous results, we have been able to run experiments where we could get the most performance out of the SISO local control system.

The controller sample rate is still 3000Hz while the control bandwidth is again limited to the 80-1000Hz frequency range. We have found that, when using a microphone located in the near field of the fans as unique reference sensor, the most attenuation at the error microphone location was achieved. Also, 256 taps are used in the control and identification filters. The error signal was differentiated and we added a feedback path removal filter. Even though these last two settings did not seem to improve the control performance, their use lead to a more stable control system.

b. Results for the Power Supply Block Fan Noise

The control results when the power supply block fan is the only operating noise source are shown in Figure 6.10 and Figure 6.11. In Figure 6.10, the most theoretically achievable attenuation (dotted red line) is compared to the achieved attenuation (dashed blue line). This attenuation represents the difference in sound pressure level at the error microphone location when the control is operating or not. We observe that the controller performance is very good at the fan BPF (250Hz) and its harmonics. On the other hand, broadband control is achieved over a limited frequency range. While the control performance target predicts broadband attenuation up to 1000Hz, the effective range is limited to about 700Hz, usually in frequency bands around the pure tones.

The same observations can be made from the results in Figure 6.11. In this plot, the resulting sound pressure level at the error microphone location is shown when the control is off (solid blue line), or operating (dashed red line). The theoretical control limit is also plotted in dotted line. The overall SPL attenuation is significant: more than 6dB achieved for 9dB predicted.

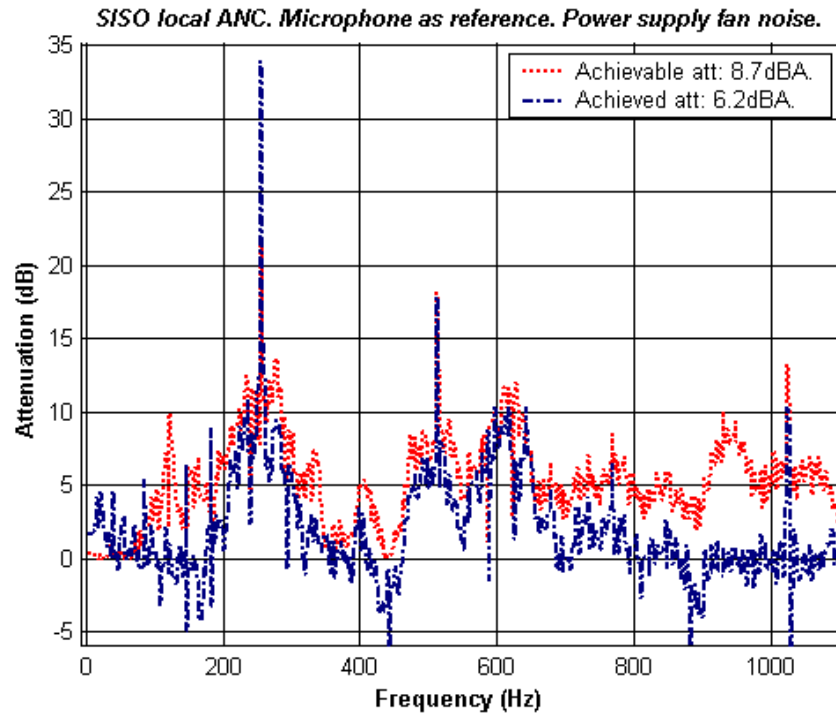


Figure 6.10 Local SISO ANC of power supply fan. Attenuation.

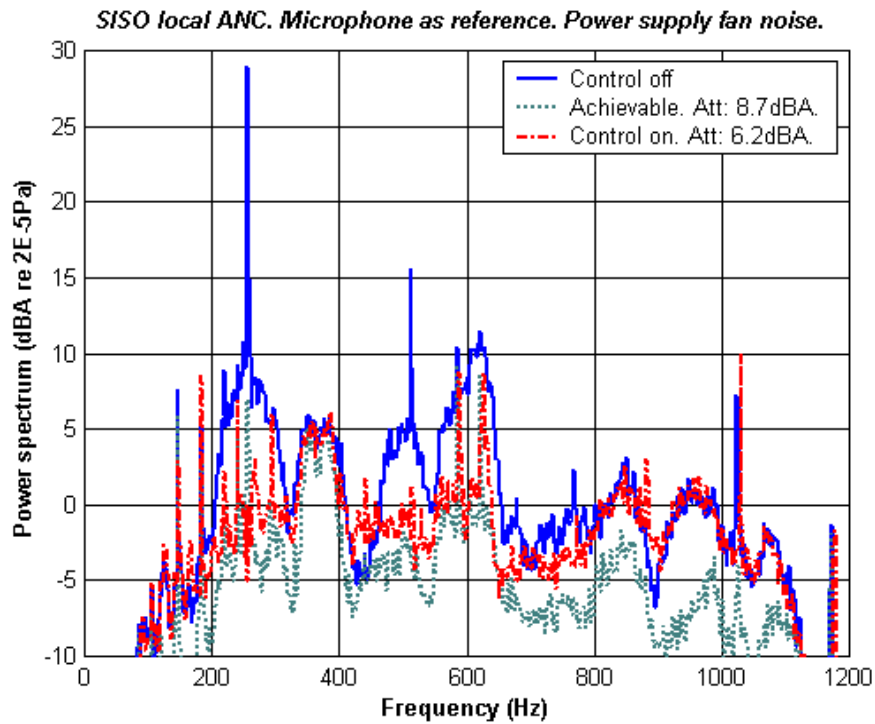


Figure 6.11 Local SISO ANC of power supply fan. Error mic power spectrum.

c. Results for the Processor Fan Noise

The control results when the processor fan is the only operating noise source are shown in Figure 6.12 and Figure 6.13. In Figure 6.12, the most theoretically achievable attenuation (dotted red line) is compared to the achieved attenuation (dashed blue line). Contrary to the power supply block fan noise, the achieved control performance is very similar to the control target. But in this case, the control seems more effective for the broadband components of the radiated noise. This phenomenon is due to the reference sensor selection. Indeed, we have seen earlier that, when using a microphone in the near field of the processor fan, only the airborne noise radiation path can be controlled. Most of the tones radiated are due to structure-borne excitation of the PC casing and are thus not controlled by this SISO controller.

The same observations can be made from the results in Figure 6.13. In this plot, the resulting sound pressure level at the error microphone location is shown when the control is off (solid blue line), or operating (dashed red line). The theoretical control limit is also plotted in dotted line. In this figure, the lack of observability (the control target is dominated by pure tones) and control of the structure-borne noise radiation path is very visible. However, the overall SPL attenuation is still quite good with more than 6dB achieved for 7dB predicted between 80 and 1000Hz.

The next step is to investigate the spatial efficiency of our local noise control system based on the size of the zone of quiet generated. The zone of quiet is determined by observing not only the SPL attenuation at the error microphone location(s), but also at monitor locations surrounding the control points. For such experiments, we will use a MIMO digital controller since it will be shown that multiple secondary noise sources and error sensors are usually needed to increase the size of the zone of quiet.

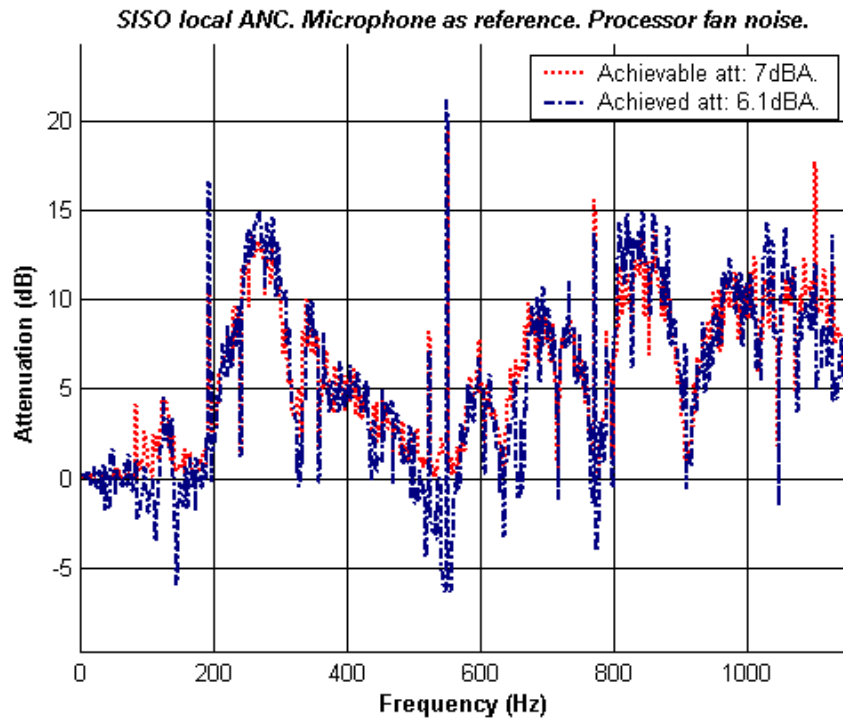


Figure 6.12 Local SISO ANC of processor fan. Attenuation.

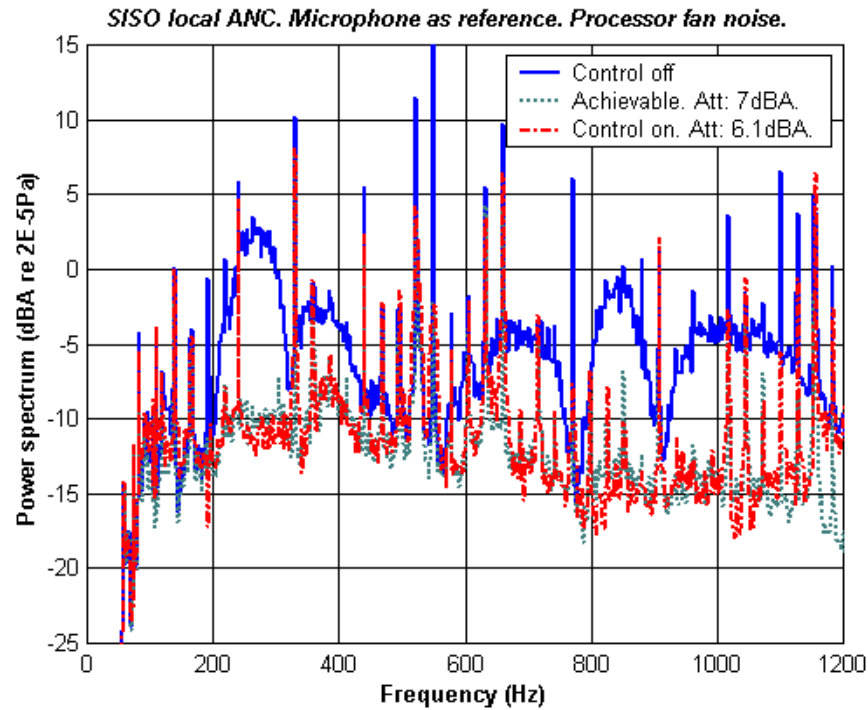


Figure 6.13 Local SISO ANC of processor fan. Error mic power spectrum.

6.3. MIMO Feedforward Control Results

6.3.1. Reference Signal Selection

As we have seen in section 6.2.1, different types of sensors and different locations, provide varied information about the primary disturbance we want to control. When studying the noise radiated from the PC, we have observed (c.f. Chapter 3) multiple radiation paths. Indeed, the primary sources studied in this project not only directly excite the surrounding acoustic space but also the structure they are connected to. Thus, the PC casing also radiates noise. The airborne paths are due to the acoustic excitation of the PC casing while the structure-borne paths are due to the mechanical excitation of the casing by the vibrating noise sources.

While the ideal reference sensor for an airborne noise radiation path is usually a microphone located in the nearfield of the source, a vibration sensor (accelerometer) will usually be the best candidate for sensing a structure-borne noise radiation path. Since both structure-borne and airborne paths are contributing to the overall noise level at the head location, the two types of reference sensors should be used together in order to perform an efficient active noise reduction.

In the previous section, a unique reference sensor, typically a microphone, was used for the separate control of each primary noise source. Fortunately, the airborne noise radiation path is a significant contributor to the noise level at the ear location. Thus, good control could be achieved even though the structure-borne path was not controlled. We will show next that even more control can be achieved when multiple reference sensors are used for each primary noise source. Moreover, when multiple independent primary noise sources are simultaneously operating, a single reference sensor is usually not sufficient to obtain information about all the disturbances. Typically, a microphone will be located in the near field of each noise source and accelerometers will be mounted on the PC casing to capture the structure borne noise contribution.

6.3.1.1. Review of the Conditioned Spectral Analysis.

a. Introduction of a New Simulation Tool

In section 6.2.1, the simple coherence function was used to select the appropriate reference sensor in the SISO control system. Unfortunately, the simple coherence function can only be used in the multiple input / single output systems if the inputs are decorrelated the one from the others. This is seldom the case in practice since, for a given set of primary disturbances, one can usually not use one reference sensor per source without picking-up contributions from the other sources. The multiple and partial coherence tools are the solution for such problem and constitute an extension of the simple coherence function to multiple inputs linear time invariant systems.

b. Tool Description

We consider a plant with multiple sources simultaneously operating. The control system is assumed to be multiple inputs (several references) single output (a unique control actuator and error sensor). Generally, we cannot predict the achievable attenuation at the error sensor by simply subtracting the correlated content of the error signal with each reference signal separately using the simple coherence function. If we do so, the attenuation will most likely be over-predicted since the reference signals are usually partially mutually correlated. The basis of multiple coherence analysis is to decorrelate the different reference signals with each other before estimating their correlated content with the error sensor. Once the input records are decorrelated, the attenuation at the error sensor corresponds to the subtraction of the correlated content of each separate input.

The process of decorrelating the multiple inputs of a MISO system is called conditioned spectral analysis and is discussed in [9]. The original set of mutually correlated inputs is transformed into a set of conditioned records that are uncorrelated with each other using the following process:

1. The simple coherence function between each input and the output signal is computed

2. The inputs are then ranked in terms of contribution to the output signal.
3. While the reference signal that is the most correlated with the output signal remains unchanged, the second most output correlated reference signal is decorrelated from the best reference. The process continues with all the remaining references, each being decorrelated from all the previous ones. Ultimately, we obtain a set of conditioned records.

There are some important conditions for the application of conditioned spectral analysis:

1. The input signals should not be strongly correlated with each other. If the simple coherence between two inputs is close to unity, then one of the correlated inputs should be suppressed since it does not bring new information about the noise source.
2. The simple coherence function between any input and the output sequence should not equal unity. If so, the system should be considered SISO because this particular input is fully coherent with the output.
3. Assuming N references are used, the next step consists of computing the multiple coherence function between any set of $N-1$ inputs and the remaining input. This function should not equal unit. If it does, it means that the remaining input is a linear combination of the other inputs and should thus be discarded.

This whole process can be effectively performed using the singular value decomposition technique. The idea is to create a matrix of coherence among all the inputs. At any given frequency, one can perform a singular value decomposition of the matrix and thus identify a subset of independent vectors (that are associated with a non-zero singular values).

Some new terms, developed in reference [9], are defined below. The partial coherence function is a simple coherence function applied to conditioned records. The multiple coherence function corresponds to the ratio of inputs-correlated output noise to the total output spectrum. It is proportional to the partial coherence function and constitutes the tool we use to predict the achievable attenuation in a MISO control system. The mathematical expressions for those tools are developed section 2.1.3.2 and originally covered in reference [9].

6.3.1.2. Application to Local MIMO ANC Control.

a. Setup

For the following experiments, the computer is placed in similar operating conditions as in section 6.2. Three cases will be studied: power supply block fan noise, processor fan noise and finally all primary noise sources operating simultaneously, including the hard disk drive. The reference signals selected various test cases are shown in Table 6.4.

Reference signal label and description :		Primary noise sources operating:		
		Power supply block fan (PS)	Processor fan (PF)	All (PS, PF, Hard Drive (HD))
a	Microphone next to the power supply fan air outlet	Yes	No	Yes
b	Microphone next to the processor fan air outlet	No	Yes	Yes
c	Accelerometer on the power supply block	Yes	No	Yes
d	Accelerometer on the processor fan	No	Yes	Yes
e	Accelerometer on the PC casing rear panel	Yes	Yes	Yes
f	Tachometer signal from the power supply block fan	Yes	No	No
g	Tachometer signal from the processor fan	No	Yes	No

Table 6.4 List of reference signals used in the MIMO local ANC system.

b. Results for the Power Supply Block Fan Noise

The achievable attenuation at a unique error sensor location, when the power supply block fan is the only operating noise source is shown in Figure 6.14. The configurations shown are:

- Solid blue line: controller not operating
- Dashed green line: controller on, reference signal ‘a’ used
- Dotted red line: controller on, reference signals ‘a’ and ‘c’ used
- Solid purple line: controller on, reference signals ‘a’, ‘c’ and ‘e’ used
- Dash-dotted gray line: controller on, reference signals ‘a’, ‘c’, ‘e’ and ‘f’ used

From the results shown in Figure 6.14, we observe that the accelerometers and tachometer do not bring much additional information about the primary noise source since the achievable attenuation is not significantly different when more than one sensor (near-field microphone) is used as reference. We conclude that the power supply block fan noise is thus mainly airborne and a microphone is sufficient to achieve good control of both broadband and pure tone noise.

In terms of overall attenuation between 200 and 1200Hz, which is the chosen frequency range of control, it seems that the larger the number of references the better. Indeed, while we can potentially achieve about 10dBA with a unique reference sensor, more than 13 dBA could be obtained with four references although this does not seem significant in Figure 6.14.

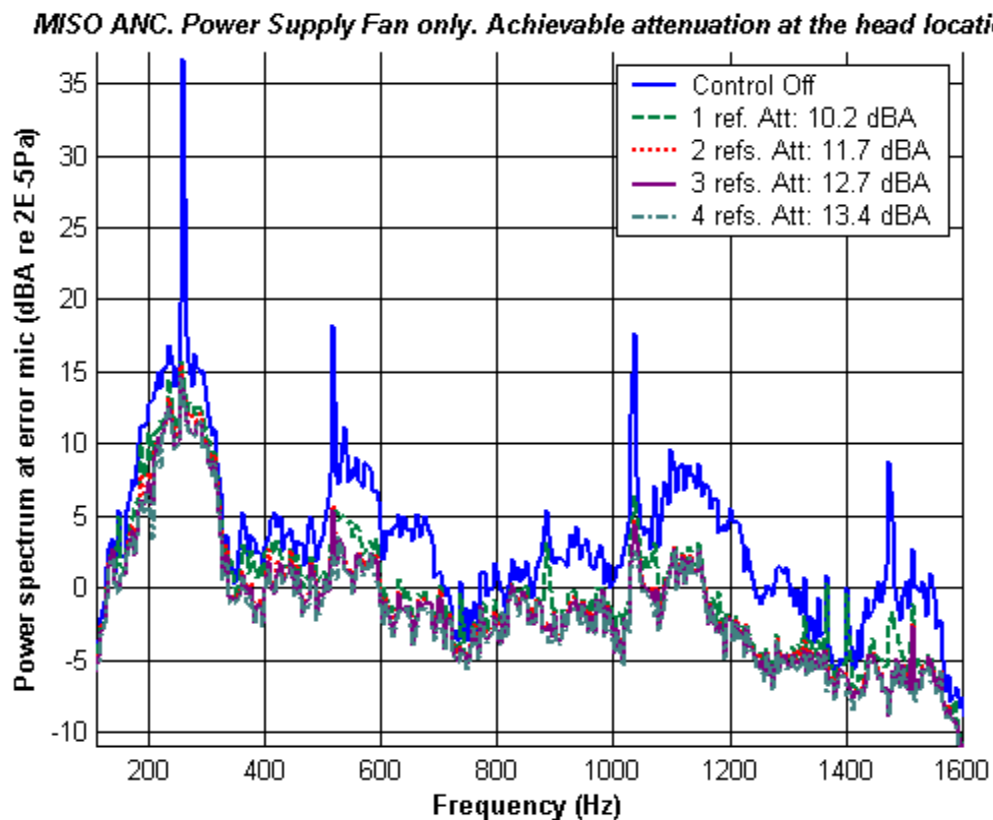


Figure 6.14 MISO reference signals selection. Power supply fan noise.

c. Results for the Processor Fan Noise

The achievable attenuation at a unique error sensor location, when the processor fan is the only operating noise source is shown in Figure 6.15. The configurations shown are:

- Solid blue line: controller not operating
- Dashed green line: controller on, reference signal ‘b’ used
- Dotted red line: controller on, reference signals ‘b’ and ‘d’ used
- Solid purple line: controller on, reference signals ‘b’, ‘d’ and ‘e’ used
- Dash-dotted gray line: controller on, reference signals ‘b’, ‘d’, ‘e’ and ‘g’ used

From the results shown in Figure 6.15, we observe that, when using only a pressure sensor (microphone near the fan outlet) as reference, only the broadband noise component at the error sensor location can be controlled. This is most likely due to the fact that the microphone located in the near field of the noise source only senses the airborne noise radiation path.

The following observations confirm this phenomenon. Indeed, when adding a single vibration sensor (accelerometer on the fan unit), the achievable control performance is greatly improved as can be seen in Figure 6.15. Comparing the dashed green line to the dotted red line, we observe that most of the pure tones are suppressed only when a vibration sensor is used. On the other hand, the addition of extras references such as a second accelerometer (on the PC casing) or a tachometer signal does not lead to significant improvements from the two references case.

Such results correlate well with what was found in section 3.2.2.2.f: the processor fan noise is a blend of airborne as well as structure-borne radiation paths. We have to point out that the fan unit was not structurally decoupled from the PC casing for these experiments, which explains why so many tones are observed between 800 and 1600Hz. As we have seen in section 5.3.2, the “high” frequency tones, that were actively controlled in the current experiments when using an accelerometer, could also be passively controlled by decoupling the processor fan from the chassis.

The passive solution for controlling the structure-borne component of the processor fan noise seem to be more appropriate than the active solution for the following reasons:

1. Although the results in Figure 6.15 show that all the tones can be actively controlled up to high frequency, such performance is not likely to be met in practice due to filter size limitations. As we have found in section 6.2.2.1.b, many “degrees of freedom” are required in the control filter for it to be able to control so many tones.
2. We will later on discuss the spatial performance of the MIMO control system. The current control results are shown at the error sensor location. As we will see, the performance typically degrades at monitor locations, the further we are from the control point. On the other hand, when structurally decoupling the fan from the PC casing, the structure-borne noise radiation path is globally controlled, at any point in space.
3. The major reason for using active noise control is that passive solutions usually require lots of space to be effective at low frequencies (see section 4.2.1.2). In the case of the structural decoupling of processor fan, we could manage to use a passive device that was very small as well as inexpensive. There is thus no apparent need for a complex active solution.

In terms of overall attenuation between 200 and 1200Hz, we also observe that two reference sensors lead to the best control improvement, i.e. from 5.5 dBA to about 8.5dBA. Then, the extra two references only help to gain 1 to 1.5 extra dBs of reduction as shown in Figure 6.15.

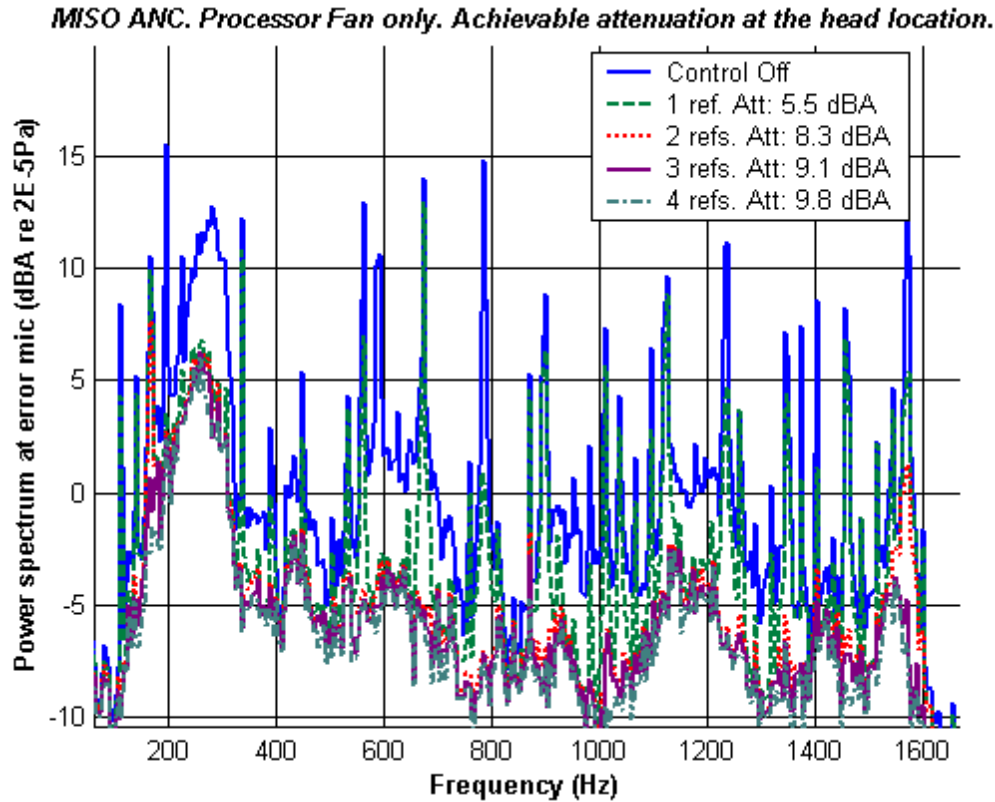


Figure 6.15 MISO reference signals selection. Processor fan noise.

d. Results With All Sources Simultaneously Operating

For the last series of experiments, all the primary sources, which are the power supply block fan, the processor fan, and the hard disk drive, are simultaneously operating. The achievable attenuation at a unique error sensor location is shown in Figure 6.16. The configurations shown are:

- Solid blue line: controller not operating
- Dashed green line: controller on, reference signal 'a' used
- Dotted red line: controller on, reference signals 'a' and 'b' used
- Solid purple line: controller on, reference signals 'a', 'b' and 'c' used
- Dash-dotted gray line: controller on, reference signals 'a', 'b', 'c', 'd' and 'e' used

From the results in Figure 6.16, we can observe the following:

- Using only a microphone in the acoustic near field of the power supply fan leads to the potential control of the power supply BPF (at 250Hz) and narrow band noise around it.
- Adding a second reference, a microphone in the acoustic near field of the processor fan, does not seem to dramatically improve the theoretical controller performance, although the overall attenuation increases from 6.4dBA to 9.3dBA.
- Adding an accelerometer to the previous set of “acoustic” reference greatly improves the achievable control with added control of many tones (fans BPF harmonics) above 400Hz.
- Using more than one extra accelerometer does not seem beneficial although the overall attenuation increases from 11.1 to 12.9dBA between the three and the five references case.

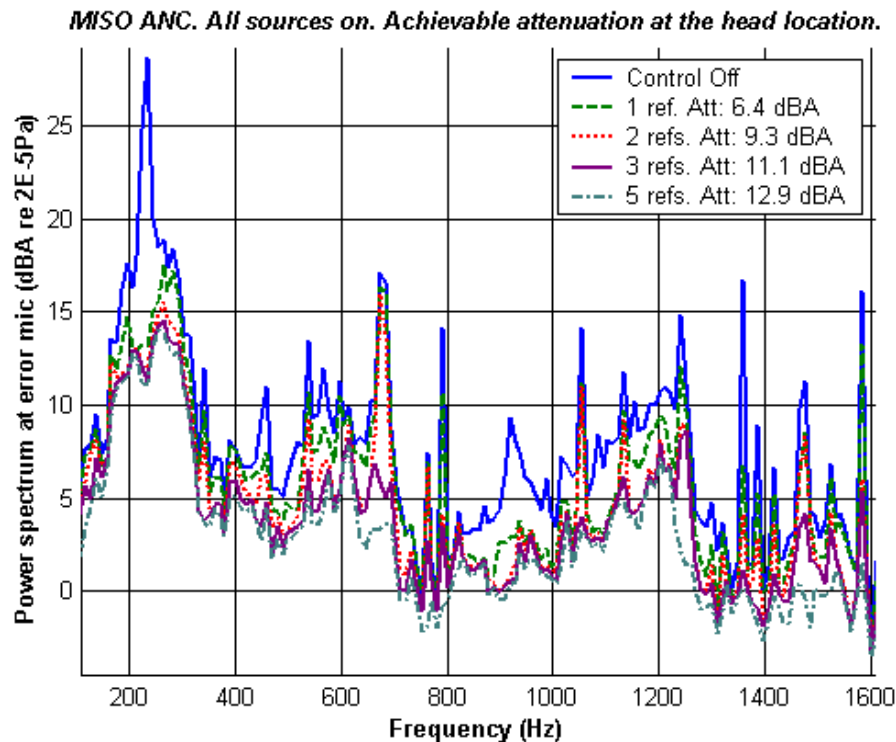


Figure 6.16 MISO reference signals selection. All sources operating.

e. Summary

From the previous experiments, we can conclude that both pressure (microphones) and vibration (accelerometer) reference sensors are required for the proper control of computer noise. The major reasons are that, not only a unique reference cannot pick-up noise of multiple simultaneously operating sources, but also some sources radiate in the surrounding acoustic space through both an airborne and a structure-borne path. We have observed that the airborne noise radiation path is effectively sensed by microphones placed in the acoustic near field of the primary source while the structure-borne path is potentially best controlled using an accelerometer in the structural near-field of the vibrating source.

The overall attenuation obtained at the error microphone location, for various combinations of primary sources and reference sensors are listed in Table 6.5. We observe a trade-off between target control performance and controller complexity (and thus implementation cost). When all the primary sources are operating simultaneously, which is the case when the PC is running normally, we believe that three reference sensors are required to achieve the most significant results. Those sensors are:

- A microphone on the rear side of the PC casing, next to the power supply fan
- A microphone on the rear side of the PC casing, next to the processor fan
- An accelerometer on the PC casing, next to the fans

Eventually, the two microphone references could be combined into a unique reference sensor. Due to time limitations, this test was not investigated. Also, an accelerometer mounted on the hard disk drive should also be part of the reference signals investigation. For the following experiments, the computer will be operating under normal conditions, that is when the three identified primary noise sources are running simultaneously. Our target will be to achieve 11dB of overall reduction at the error sensor location using three reference signals (two microphones and one accelerometer).

Achievable attenuation at the error microphone in dBA (in the 200-1200Hz frequency range)					
<i>Number of reference signals used:</i>	<i>1</i>	<i>2</i>	<i>3</i>	<i>4</i>	<i>5</i>
Power supply block fan noise:	10.2	11.7	12.7	13.4	NA
Processor fan noise:	5.5	8.3	9.1	9.8	NA
All sources operating (including hard drive):	6.4	9.3	11.1	12.0	12.9

Table 6.5 Multiple reference signals selection for the MISO local ANC.

6.3.2. Control Results

6.3.2.1. Influence of the Number of Reference.

a. Control Setup

For the following experiments, we operate the PC in standard conditions (the simulated user-environment is shown in Figure 6.2) with the power supply block fan, processor fan and hard disk drive simultaneously operating. A single secondary noise source (active) and a unique error microphone (B&K type 3166) at the head location of the operator are used. The three reference sensors used are a microphone next to the air outlet of power supply fan, a microphone next to the air outlet of the processor fan and an accelerometer on the PC casing (rear panel, where the processor fan is attached).

These ANC experiments take advantage of the causality study carried out for the Global ANC (c.f. section 5.3.1.3). Even though causality is not as important an issue in the present case (the primary path propagation delay is much larger than in the ducts), we make sure it will not be a problem by adequately filtering the different signals and sampling the controller at the fastest rate possible.

The controller settings are listed in Table 6.6. We can observe that, even though the control bandwidth is limited to the 20-1250Hz frequency range, the control sample rate is higher than twice the higher frequency to control (i.e. 1250Hz). This choice is made to minimize the computation delay in the digital controller. Moreover, the reference and control signals are low-pass filtered at 3150 and 1600Hz respectively. These values are deliberately higher than the cut-off frequency of the low-pass filter used for the error signal. The reason is that it was found in section 5.3.1.3 that, the higher the low-pass filter cut-off frequency, the smaller the Group Delay through the filter. We will note however that the reconstruction filter cut-off frequency is above half the controller sample rate. Thus, we might observe some spillover during the control phase, which is characterized by an increase in sound pressure level at some frequencies, outside the frequency range of control.

The controller used for the following experiments is the same controller as used for the Global ANC final test. This feedforward adaptive digital controller is based on the multiple channel Filtered-X LMS algorithm. The hardware used is a Texas Instruments C40 series DSP board interfaced on a host PC using a Labview based program developed in the VAL.

<i>Control system settings:</i>	<i>Value / Range:</i>
Reference signals analog filter:	Lowpass 3150Hz
Control signal analog filter:	Lowpass 1600Hz
Error signal analog filter:	Bandpass 20-1250Hz
Controller sample rate:	5000Hz
Digital identification filter:	120 taps
Digital control filter:	120 taps
Feedback removal filter	No
Error signal differentiation:	Yes

Table 6.6 MISO controller settings for the reference signals tests

b. Results

The MISO controller performance is shown in Figure 6.17. The plots, from top to bottom respectively, show the achieved control performance at the head location of the operator when one, two, or three reference sensors are used in the controller. For the three graphs, the curves are the following:

- Blue dotted line: controller not operating
- Green dash-dotted line: most theoretically achievable control
- Red solid line: achieved control performance when the control is operating

We observe in Figure 6.17 that the power supply fan BPF is usually well controlled in all three configurations. The use of several reference sensors seems beneficial to control more of the broadband noise component. On the other hand, regardless of the number of reference sensors used, the effective controller is not able to control most of the “high-frequency” pure tones radiated by the various PC noise sources. In particular, we observe in the lowermost plot of Figure 6.17 that the lack of coherence is not the reason why all these tones are not controlled. We believe that the problem here is the filter size. Indeed, only 120 taps are used for both the control and system ID filters. This might not be sufficient to control both the broadband as well as multiple tones contributions of the primary noise sources.

The overall attenuation results are listed in Table 6.7. It seems clear that the filter size is a limiting factor for the control performance. However, simply controlling the power supply fan noise leads to 6dBA of reduction, which is a good starting point.

<i>Number of references used:</i>	<i>Attenuation (dBA 100-1250Hz frequency range)</i>	
	Predicted	Achieved
1	6.9	5.8
2	9.9	7.4
3	11.0	8.3

Table 6.7 Reference signals effect on MISO ANC. All sources operating.

MISO local ANC. All sources on. Reference signals effect on noise level at head location.

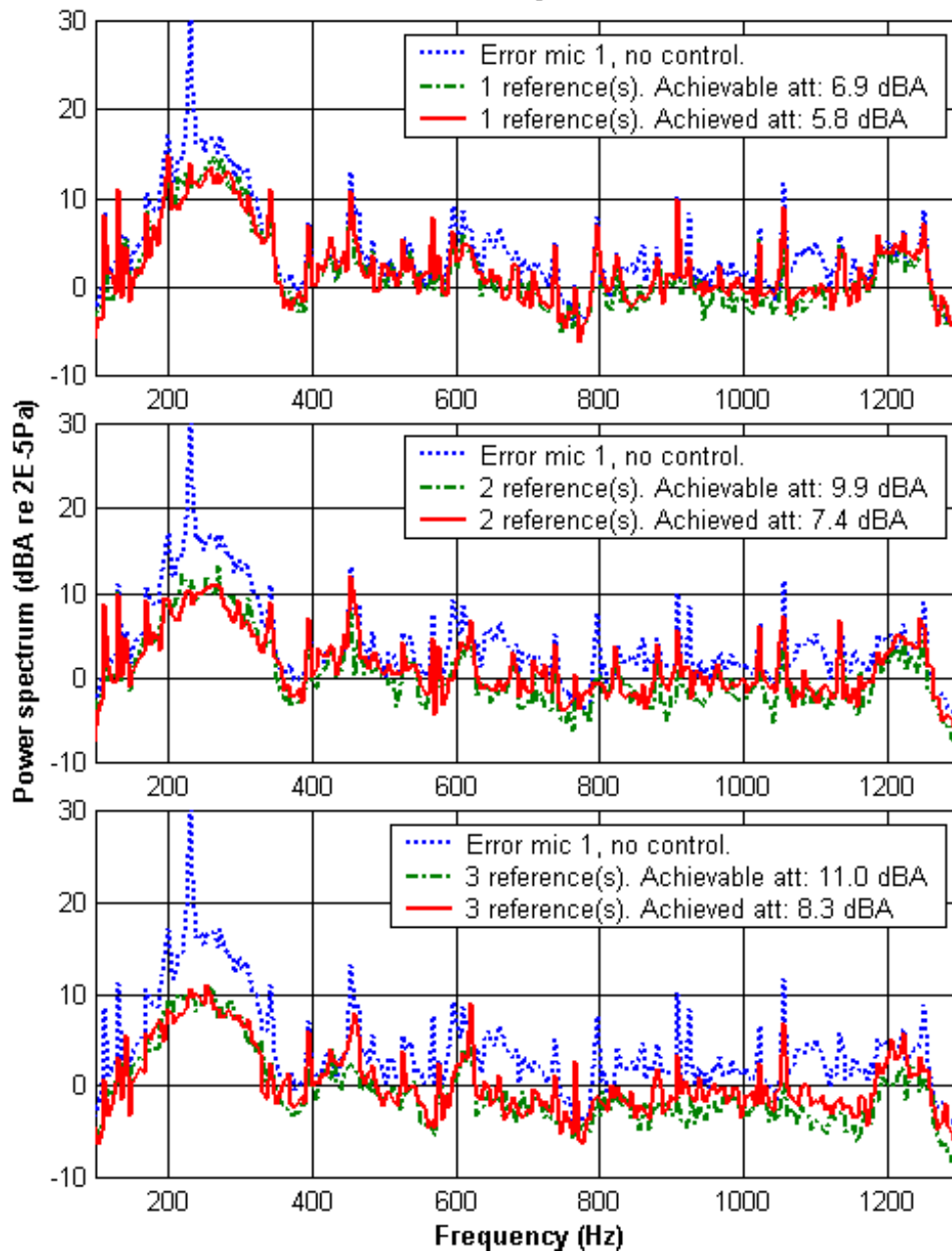


Figure 6.17 Reference signals effect on MISO ANC. All sources operating.

6.3.2.2. Effect of the Source Complexity on the Performance.

a. Test Setup

The objective of this test is to determine the influence of the number of primary noise sources operating on the performance of the noise controller. The control setup and controller settings are identical to the tests performed in section 6.3.1. Four different configurations are tested involving the used of two to three reference sensors. A test setup summary is shown in Table 6.8.

The hard drive noise will not be a noise source actively controlled due to its noise characteristics, which are above any practical range of Active Noise Control (see section 3.2.2.3). Nevertheless, its operation can eventually affect the performance of the ANC system dedicated to the fans noise. Thus, one the following tests includes the hard disk drive as operating noise source. However, it is important to note that we have structurally decoupled the hard drive from the PC casing for the following experiments. For more information regarding the passive control of hard drive noise, one should refer to section 1.1.

Reference signal used:	<i>Simultaneously operating sources</i> (<i>PS=Power Supply Fan / PF=Processor Fan / HD=Hard Drive</i>)			
	<i>PS only</i>	<i>PF only</i>	<i>PS + PF</i>	<i>PS + PF + HD</i>
Microphone near PS	<i>Yes</i>	<i>No</i>	<i>Yes</i>	<i>Yes</i>
Microphone near PF	<i>No</i>	<i>Yes</i>	<i>Yes</i>	<i>Yes</i>
Accelerometer on PC casing	<i>Yes</i>	<i>Yes</i>	<i>Yes</i>	<i>Yes</i>

Table 6.8 Noise source complexity effect MISO ANC. Configurations tested.

b. Control Results at the Error Microphone Location

The results of the tests are shown in Table 6.9 and Figure 6.18. The blue dotted line in Figure 6.18 corresponds to the sound pressure level at the error microphone when the controller is off, while the solid red line corresponds to the achieved control performance using a single speaker. The four plots in Figure 6.18 correspond to:

- Top left graph: Power supply block fan operating
- Top right graph: Processor fan operating
- Bottom left graph: Power supply and processor fans operating
- Bottom right graph: All sources simultaneously operating

We observe that the best control performance is achieved when the power supply block fan is the only operating noise source (top left graph) with 11dBA of overall SPL reduction at the operator head location between 100 and 1250Hz. However, this result is mainly due to the shape of the spectrum content. It is dominated by the fan BPF at 250Hz, which is 20 to 30dB above the background noise.

When the processor fan is the only operating noise source (top right graph), control of both the multiple tones (structure-borne path) and broadband noise (airborne path) is good. This is probably due to the fact that two reference sensors are used. The overall attenuation is not as significant as for the power supply fan noise with 7.1dBA across the frequency range of interest.

When both the power supply and processor fans are simultaneously operating (bottom left graph), there is limited control of the broadband noise component while most of the tones (except at 500Hz) are fully controlled. The overall performance, 7.6dBA, is close to what was achieved with only the processor fan running.

Finally, the addition of the hard disk drive noise (bottom right graph) does not seem to affect the controller performance. Actually, the overall performance, 8.7dBA, is even higher than the previous case. However, this might be due to variations in sources characteristics from one test to the other, which in turns could affect the control performance.

<i>Operating sources:</i>	<i>PS only</i>	<i>PF only</i>	<i>PS + PF</i>	<i>PS + PF + HD</i>
Sound Pressure Level (Control Off, dBA):	37.7	28.5	35.1	35.9
Achieved attenuation (100-1250Hz, dBA):	11.0	7.2	7.6	8.8

Table 6.9 Noise source complexity effect MISO ANC. Results.

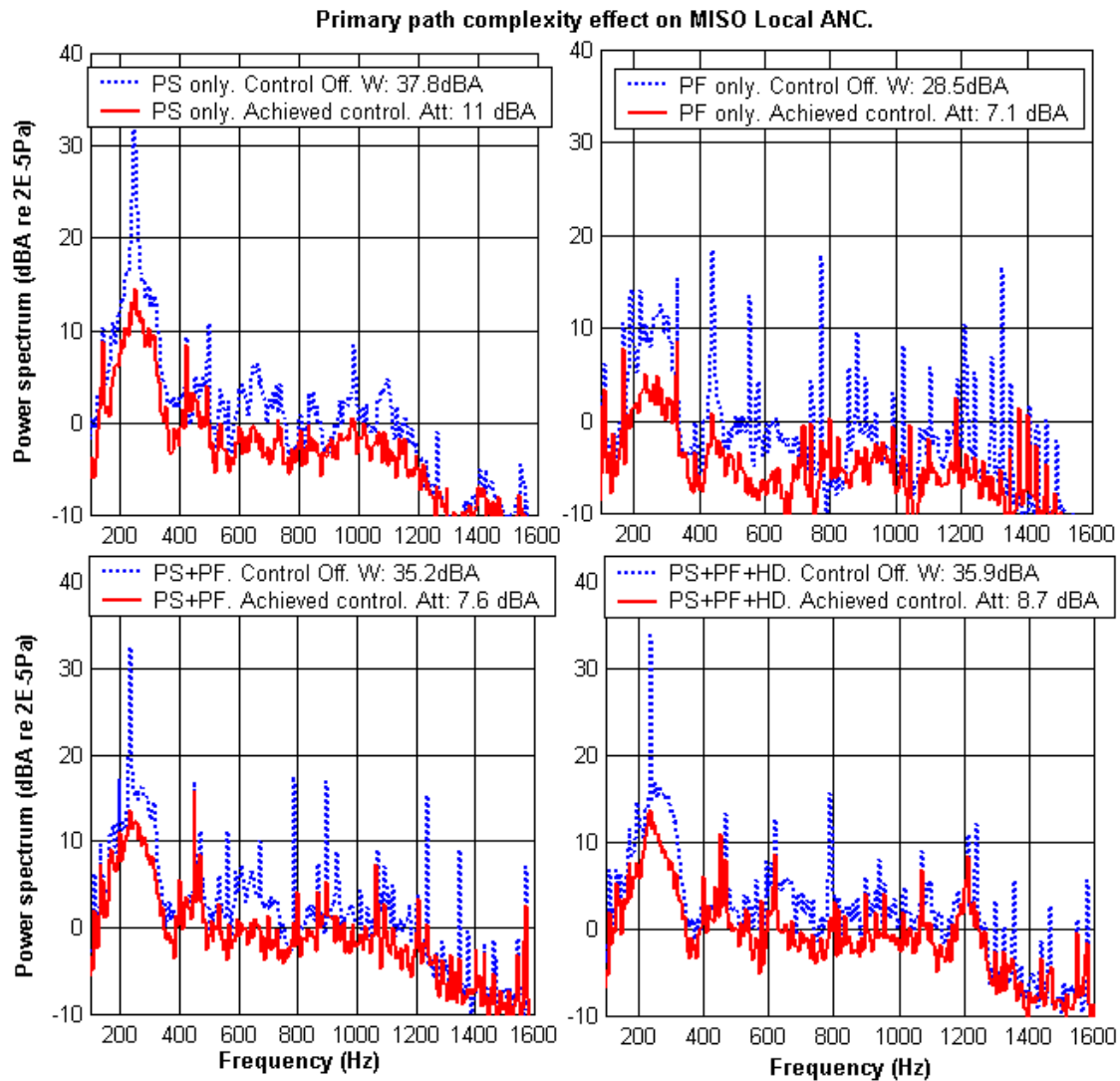


Figure 6.18 MISO Local ANC performance. Effect of primary noise sources complexity.

6.3.2.3. Description of the Zone of Quiet.

a. Introduction

So far, we have only examined the control performance at the error sensor location. We will explain below that, in a local noise control strategy, the control performance may vary depending on where the performance is measured in space. The basis principle for this phenomenon is the interference between the primary and secondary sound fields. When the control performance is measured not only at the error sensor location but also at surrounding monitor sensor locations, one is able to plot the spatial extent of the control performance. The volume in which the SPL attenuation is higher than a given threshold value is then called the zone of quiet.

b. Interference Phenomenon for an Acoustic Dipole

Dipole Formulation:

We now consider a simple analysis to describe the behavior of source interference. For simplification purposes, we assume the primary and secondary noise sources have the radiation characteristics of a monopole. For a detailed description of the phenomenon, one should refer to [9]. The complex pressure radiated by a point source can be expressed in spherical coordinates by the following [9]:

$$p(r) = \frac{j \cdot \omega \cdot \rho_0 \cdot q \cdot e^{-j \cdot k \cdot r}}{4 \cdot \pi \cdot r} \quad (6.1)$$

In equation (6.1), q represents the complex source strength and it has the dimension of a volume velocity. We can now assume a sound field generated by a primary noise source having the source strength q_p and a secondary point monopole source of source strength q_s , which are respectively at the distance r_p and r_s from an observation point $p(r, \theta)$ as shown in Figure 6.19. The pressure can now be expressed as:

$$p(r, \theta) = \frac{j \cdot \omega \cdot \rho_0 \cdot q_p \cdot e^{-j \cdot k \cdot r_p}}{4 \cdot \pi \cdot r_p} + \frac{j \cdot \omega \cdot \rho_0 \cdot q_s \cdot e^{-j \cdot k \cdot r_s}}{4 \cdot \pi \cdot r_s} \quad (6.2)$$

The monopoles are separated by the distance d . The defined distance r and angle θ are illustrated in Figure 6.19.

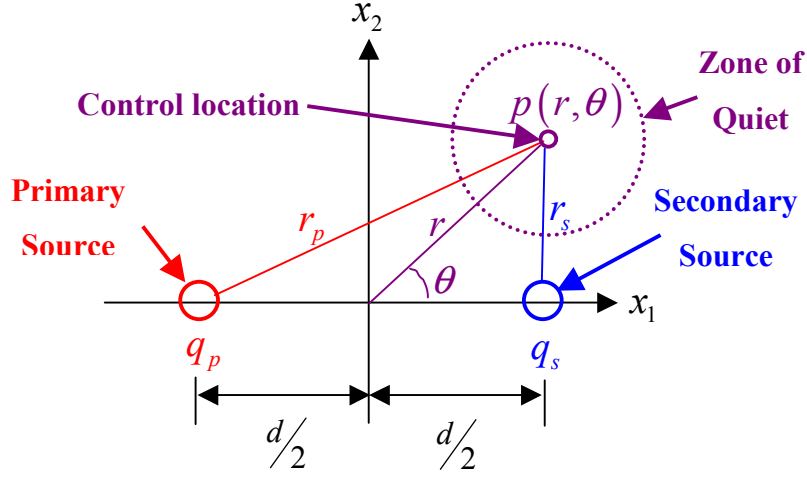


Figure 6.19 Primary and secondary monopole sources, separated by distance d .

We consider that the observation point is in the far field of the sources:

$$k \cdot r = \frac{2 \cdot \pi}{\lambda} \cdot r = \frac{2 \cdot \pi \cdot f}{c} \cdot r \gg 1, \quad (6.3)$$

where:

k is the acoustic wavenumber in m^{-1}

λ is the acoustic wavelength in m

f is the frequency in Hz

c is the sound speed in m/s

In such case, we can make the approximations that:

$$r_p \approx r_s \approx r \quad \text{for the denominator of equation (6.2)} \quad (6.4)$$

$$r_p \approx r + \left(\frac{d}{2}\right) \cdot \cos(\theta) \quad \text{for the numerator of equation (6.2)} \quad (6.5)$$

$$r_s \approx r - \left(\frac{d}{2}\right) \cdot \cos(\theta) \quad \text{for the numerator of equation (6.2)} \quad (6.6)$$

Then, equation (6.2) becomes:

$$p(r, \theta) = \frac{j \cdot \omega \cdot \rho_0}{4 \cdot \pi \cdot r} \cdot \left[q_p \cdot e^{-j \cdot k \cdot \left(r + \frac{d}{2} \cdot \cos(\theta)\right)} + q_s \cdot e^{-j \cdot k \cdot \left(r - \frac{d}{2} \cdot \cos(\theta)\right)} \right] \quad (6.7)$$

Acoustic Doublet:

We first consider that the two monopoles sources have equal strengths, but are 180 degrees out of phase with each other, i.e. $q_p = -q_s = q$. By substituting those values in equation (6.7), we obtain the resulting sound pressure in the far field:

$$p(r, \theta) = 2 \cdot \sin\left(\frac{kd}{2} \cdot \cos(\theta)\right) \cdot \frac{\omega \cdot \rho_0 \cdot q}{4 \cdot \pi \cdot r} \cdot e^{-j \cdot k \cdot r} \quad (6.8)$$

For frequencies such that $kd \ll 1$, this equation further reduces to:

$$p(r, \theta) = kd \cdot \cos(\theta) \cdot \frac{\omega \cdot \rho_0 \cdot q}{4 \cdot \pi \cdot r} \cdot e^{-j \cdot k \cdot r} \quad (6.9)$$

The pressure amplitude distribution for the sources combination is:

$$|p(r = r_0, \theta)| = kd \cdot \frac{\omega \cdot \rho_0 \cdot q}{4 \cdot \pi \cdot r_0} \cdot |\cos(\theta)| \quad (6.10)$$

In such configuration, the pressure amplitude decreases as $1/r$, like the monopole source. But the important aspect is that, for a fixed r , it has a dependence on θ as shown in Figure 6.20. We can observe that the interference between the primary and secondary sources creates a zone of quiet where the pressure amplitude is zero.

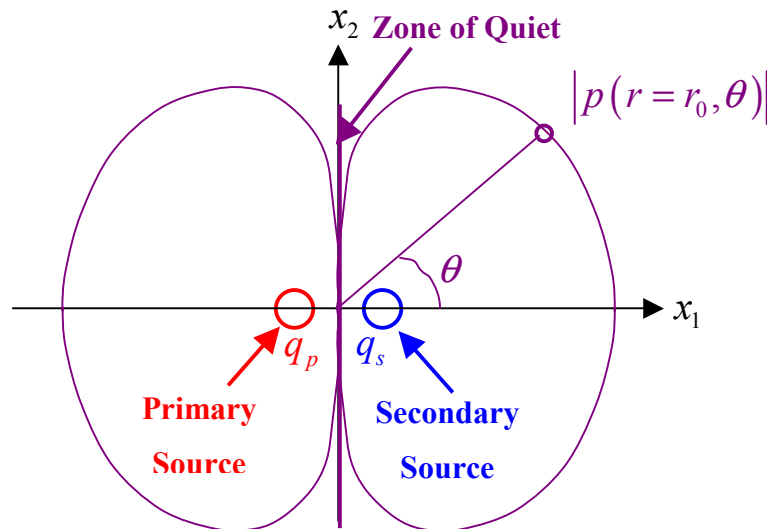


Figure 6.20 Zone of quiet for an “acoustic doublet” [26].

Sound Pressure Cancellation at a Specific Location:

We now adjust the secondary source strength in order to cancel the pressure amplitude at the observation point defined in Figure 6.19. From the expression in equation (6.7), we can deduce q_s such the far field pressure is zero at any distance r and angle $\theta = \theta_1$:

$$p(r, \theta_1) = 0 = \frac{j \cdot \omega \cdot \rho_0}{4 \cdot \pi \cdot r} \cdot \left[q_p \cdot e^{-j \cdot k \cdot \left(r + \frac{d}{2} \cdot \cos(\theta_1) \right)} + q_s \cdot e^{-j \cdot k \cdot \left(r - \frac{d}{2} \cdot \cos(\theta_1) \right)} \right] \quad (6.11)$$

Then:
$$q_s = -q_p \cdot e^{-j \cdot k \cdot d \cdot \cos(\theta_1)} \quad (6.12)$$

At any angle θ and distance r , equation (6.11) becomes:

$$p(r, \theta) = \frac{j \cdot \omega \cdot \rho_0}{4 \cdot \pi \cdot r} \cdot q_p \left[e^{-j \cdot k \cdot \left(r + \frac{d}{2} \cdot \cos(\theta) \right)} - e^{-j \cdot k \cdot \left(r - \frac{d}{2} \cdot \cos(\theta) + d \cdot \cos(\theta_1) \right)} \right] \quad (6.13)$$

From the expression above, it can be clearly seen that the noise cancellation will be perfect only at angles $\theta = \theta_1$ modulo π . A visual description of this interference phenomenon is shown in Figure 6.21. At some angles, the sound pressure level will even be higher than if only the primary source was operating.

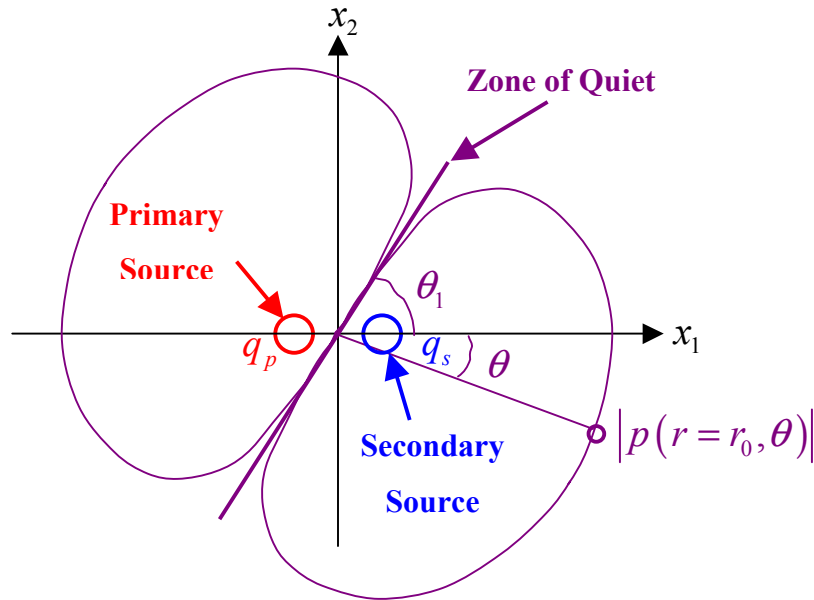


Figure 6.21 Zone of quiet for a “tuned acoustic doublet” [26].

Attenuation of the Primary Sound Field at all Angles:

We now adjust the secondary source strength such that the sound pressure level at any distance r (in the far field) and any angle θ is small in the case of a dipole compared to the primary source operating alone. First, we express equation (6.13) as a function of the complex pressure $p(r, \theta)$ due exclusively to the primary source $p_p(r, \theta)$:

$$p(r, \theta) = p_p(r, \theta) \cdot \left[1 - e^{j \cdot k \cdot d \cdot (\cos(\theta) + \cos(\theta_1))} \right] \quad (6.14)$$

Then, using the identity $|1 - e^{-j \cdot x}|^2 = 2 \cdot (1 - \cos(x))$, we can compare the modulus squared of the far field complex pressure produced by the interference of the two sources to that produced by the primary source alone [9]:

$$\frac{|p(r, \theta)|^2}{|p_p(r, \theta)|^2} = 2 \cdot \left[1 - \cos(kd \cdot [\cos(\theta_1) - \cos(\theta)]) \right] \quad (6.15)$$

It follows that a simple condition for producing attenuation of the primary field at all angles θ , regardless of the choice of θ_1 is:

$$|p(r, \theta)|^2 < |p_p(r, \theta)|^2 \text{ if } [1 - \cos(2 \cdot k \cdot d)] < \frac{1}{2} \quad (6.16)$$

In other words, provided that the primary and secondary noise sources are separated by less than one-twelfth of the wavelength of the sound being radiated ($d < \lambda/12$), attenuation of the primary sound field will then be global. At the particular angle $\theta = \theta_1$, there will even be totally destructive interference. The condition for global reduction of the sound field is frequency dependent as shown in equation (6.17).

$$d < \lambda/12 = \frac{c}{12 \cdot f} \quad (6.17)$$

The source separation distance criterion described above for producing global reductions in sound pressure in the far field can be less stringent if we do not force the sound pressure to be zero at a particular angle.

If we denote the following:

- p_{pp} the pressure at the position of the primary source due to the primary source
- p_{sp} the pressure at the position of the primary source due to the secondary source
- p_{ps} the pressure at the position of the secondary source due to the primary source
- p_{ss} the pressure at the position of the secondary source due to the secondary source

Then, the total source power output can be written as [9]:

$$W = \frac{1}{2} \cdot \text{Re}\{p^* \cdot q\} = \frac{1}{2} \cdot \text{Re}\{(p_{pp} + p_{sp})^* \cdot q_p\} + \frac{1}{2} \cdot \text{Re}\{(p_{ps} + p_{ss})^* \cdot q_s\} \quad (6.18)$$

The previous expression can be reduced to the complex quadratic form:

$$W = \frac{\omega^2 \cdot \rho_0}{8 \cdot \pi \cdot c_0} \cdot (|q_s|^2 + q_s^* \cdot a + a^* \cdot q_s + b), \quad (6.19)$$

where:

$$a = q_p \cdot \frac{\sin(k \cdot d)}{k \cdot d}$$

$$b = |q_p|^2$$

This quadratic function has a unique minimum [9] specified by the optimal secondary source q_{s0} leading to the minimum power output W_0 :

$$q_{s0} = -q_p \cdot \frac{\sin(k \cdot d)}{k \cdot d} \quad (6.20)$$

$$W_0 = W_{pp} \cdot \left[1 - \left(\frac{\sin(k \cdot d)}{k \cdot d} \right)^2 \right] \quad (6.21)$$

The previous expressions lead to an important conclusion. As the separation distance between the two sources increases, the secondary source strength asymptotes to zero. In particular, once the separation distance exceeds $\lambda/2$, little can be done to reduce the net power radiated to the far field by the primary source.

If the sources become separated by a distance that is large compared to the acoustic wavelength, then the control at a particular point in space must be achieved at the expense of a net increase in the total acoustic energy radiated [9]. Thus, if we define the zone of quiet as the locations in space where the sound field is attenuated, we can observe from the simple dipole example that it is a function of the sources spacing and frequency. This is one of the reasons why do not attempt to perform local active noise control above 1000 to 1500Hz, which corresponds to an acoustic wavelength of approximately 22 to 35 cm, since we can rarely collocate the primary and secondary noise sources with such “accuracy”.

Eventually, we are interested in a different definition of the zone of quiet. In the local active noise control strategy, the zone of quiet can be defined as the domain in space, around the error sensor, where the sound pressure is attenuated. In a local noise control strategy, where the objective is to minimize the mean square sound pressure level at a given set of discrete locations in space, it is usual that the optimization process be performed at the expense of an increase in SPL at uncontrolled locations for the reasons described earlier. We will not be able to illustrate in details how the “local zone of quiet” is affected by the source configuration. However, we can postulate that, similarly to what was observed earlier, its size will likely reduce as the frequency (or rather the ratio between acoustic wavelength and sources separation distance) increases. The objective of the experiments in the next section is to investigate the impact of the primary and secondary sound fields complexity as well as the number of error sensor locations on the spatial extent of the control.

6.3.2.4. Effect of the Source Complexity on the Spatial Control Performance.

a. Control Setup

The following experiments aim at showing the effect of the primary sound field complexity on the size of the generated zone of quiet. For the first time, we use multiple secondary noise sources and error microphones (two speakers on the desk combined with two error microphones located approximately at the left and right ear location of a PC user in a typical position). The only other difference with the previous local active noise control experiments is that the frequency range of control extended up to 1600Hz.

In Figure 6.22, we show in a picture of the PC in a simulated user environment, with the two control sources, as well as the two B&K ½" microphones used as error sensors at the location of the operator's ears. The computer is thus located in a VAL anechooic chamber while a floor, one side and a back wall are simulated using hard wood boards. The speakers used are 5" diameter drivers in a sealed enclosure of small volume.

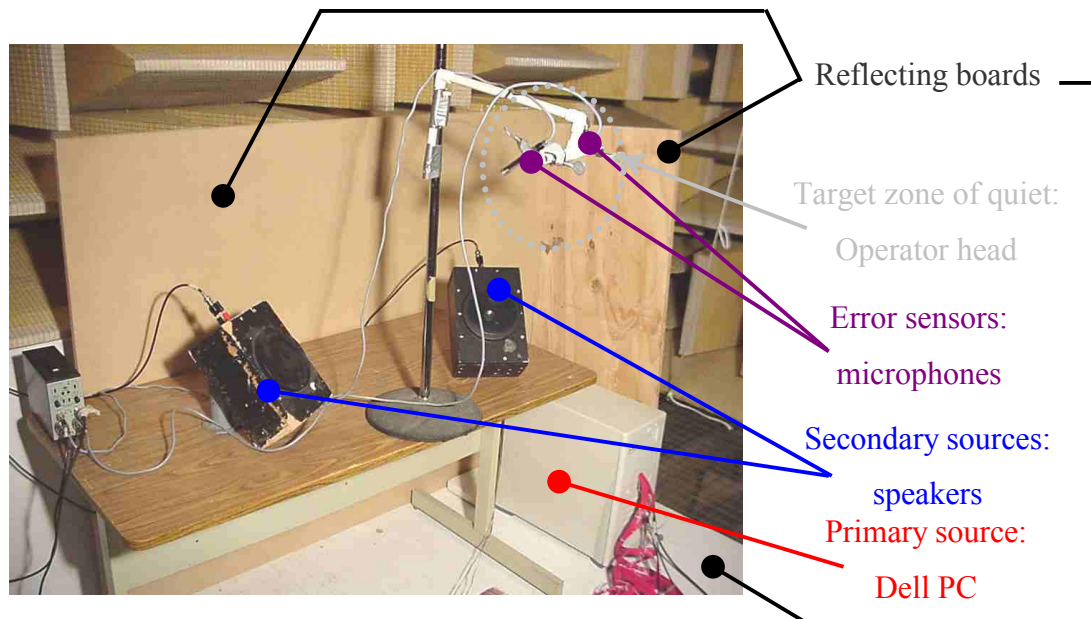


Figure 6.22 Picture the MIMO Local ANC setup in the simulated user environment.

The computers used for the data acquisition, active noise control, the analog filters, signal conditioners and speaker amplifier are shown in Figure 6.23. This equipment is located outside the anechoic chamber in order to keep the background noise to a low level inside the chamber.

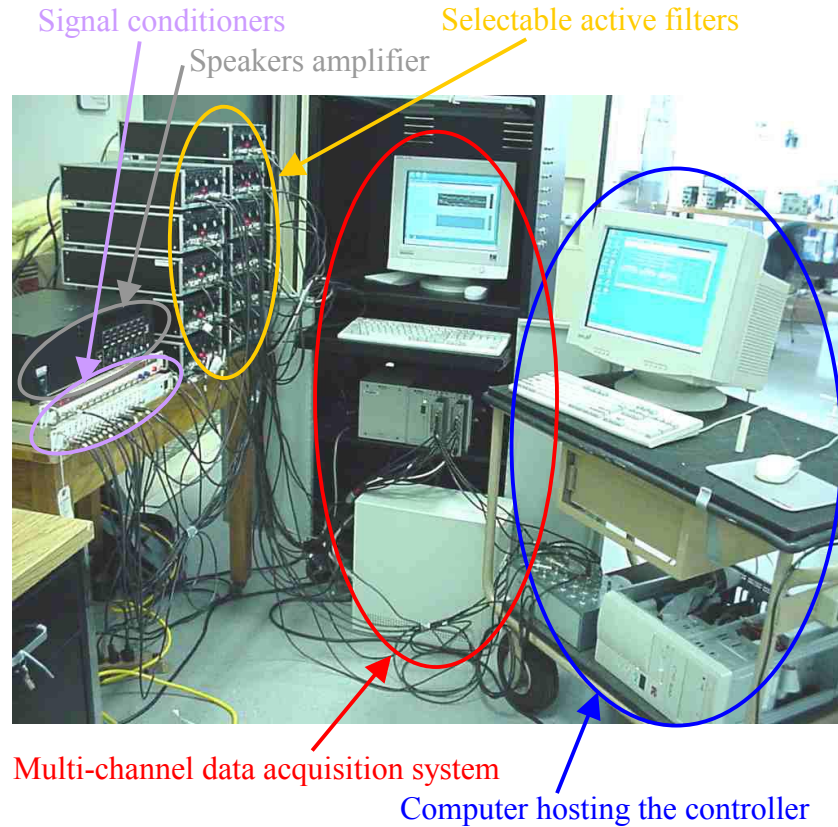


Figure 6.23 Picture the MIMO Local ANC hardware setup.

b. Description of the Monitoring System

In order to monitor the pressure at multiple locations around the error sensors, we built a microphones array using 1/4" piezo type PCB sensors. The monitor microphones spacing was 6 inches in all directions as shown in Figure 6.24, which corresponds to the average spacing between human ears. A total of 36 monitor locations were thus recorded. Due to limitations in the number of sensors available, the operation was performed in three steps, simultaneously recording the twelve microphones in one horizontal plane.

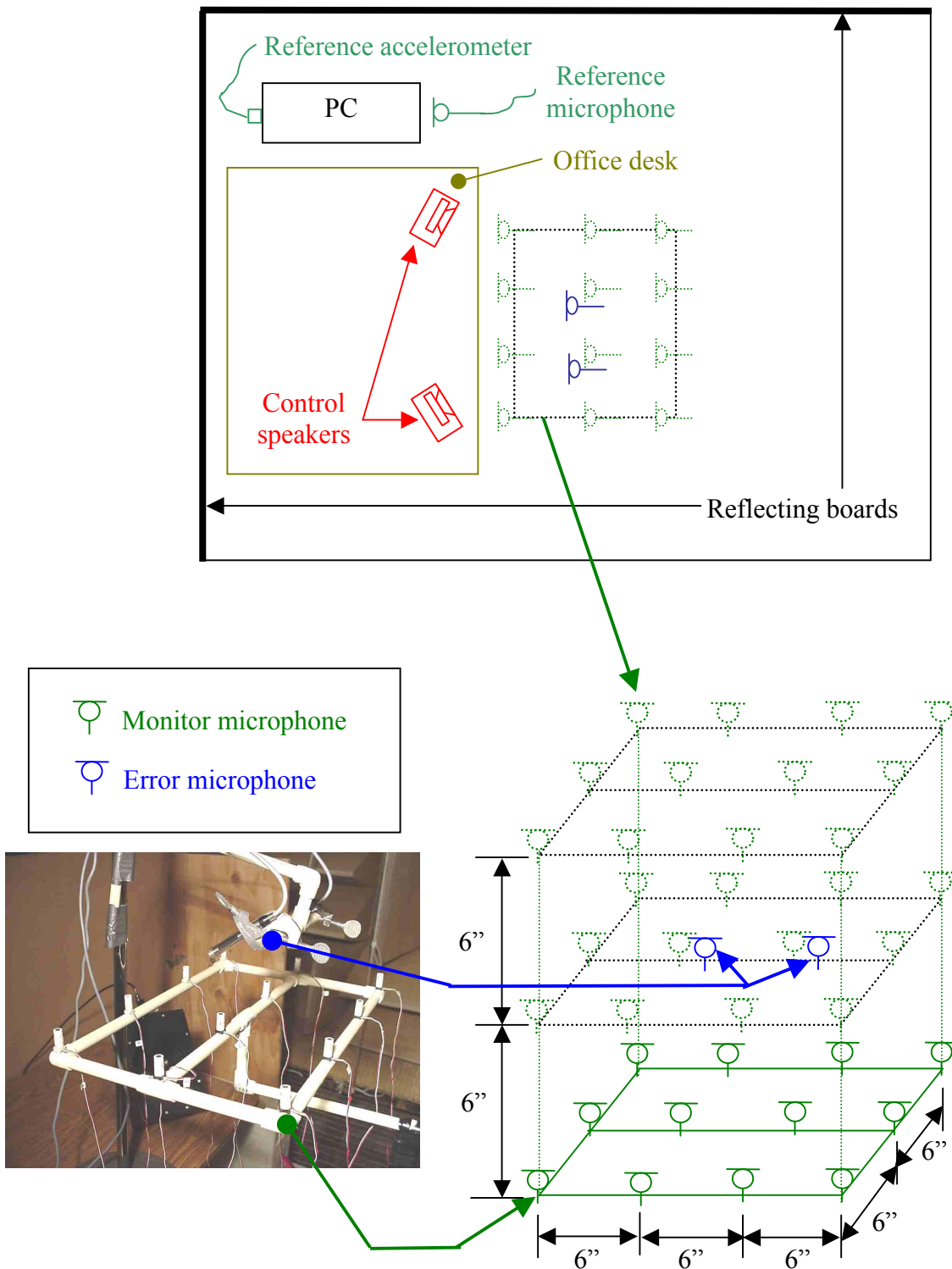


Figure 6.24 MIMO local ANC setup with the monitoring microphones array.

*c. Results**Overall attenuation:*

The monitoring microphones shown in Figure 6.24 were used to generate 3-dimensional plots of the attenuation around the user head location. Figure 6.25 through Figure 6.28 show the attenuation achieved at the user head location for the four following sources configuration:

- Figure 6.25: Power supply block fan operating
- Figure 6.26: Processor fan operating
- Figure 6.27: Power supply block fan and processor fan operating
- Figure 6.28: Power supply block fan, processor fan and hard drive operating (hard drive structurally decoupled from the PC casing).

The pressure measurements recorded at the 36 monitoring locations are interpolated in various planes order to form a smooth 3-dimensional rendering of the sound field. In Figure 6.25 through Figure 6.28, the value computed at each monitoring location is the overall sound pressure attenuation over the 100-1250Hz frequency range in dBA. The attenuation level is rendered using a “temperature” color map. Thus, blue or “cold” regions represents zone of small attenuation while the red or “hot” regions (typically in the vicinity of the error sensors) are zones where the attenuation is large. In the following figures, we also plot a virtual head in dotted lines, which represents the typical operator’s head.

We start the analysis with the attenuation results when the power supply block fan is the only operating noise source (c.f. Figure 6.25). The maximum attenuation (about 7dBA) is achieved at the error microphones locations. The control performance degrades as the distance to the error microphones decreases, which creates approximately spherical surfaces of equal attenuation centered between the two error sensors.

However, it seems that the control performance in the axis of the secondary sources is good, regardless of the distance to the error sensors. This result might suggest that a

“cone” of silence, whose solid angle is not well known, is generated by the secondary sources. Since the primary and secondary sources are mostly radiating in a free field environment, this phenomenon is possible. The analogy with the dipole sound source is that the sound field could be cancelled at a particular angle, at any distance in the far field of the sources. It is also important to observe that there is no amplification of the sound field at any of the monitoring locations.

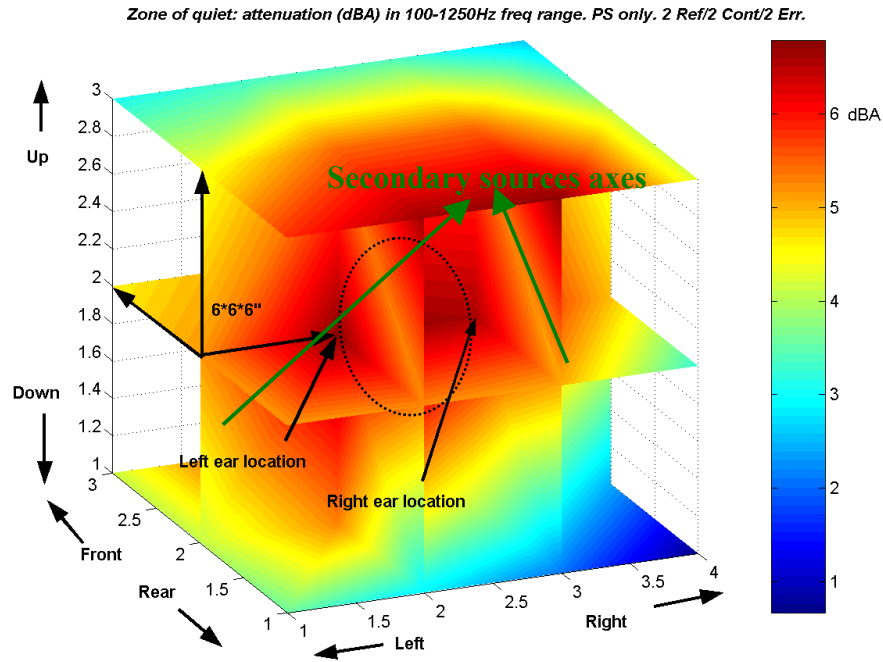


Figure 6.25 Spatial control performance. Power supply fan noise.

Now, we look at the control results when the processor fan is the only operating noise source (c.f. Figure 6.26). The maximum performance is now only 3dBA next to the error sensors. Moreover, compared to Figure 6.25, the size of the zone of quiet seems smaller and its shape is rather oval. The reason why the size of the zone of quiet is smaller in the present case will be later investigated. We believe the differences in size is due to the fact that the spectral content of processor fan noise is higher frequency than that of the power supply fan, which will inevitably affect the spatial extent of the control.

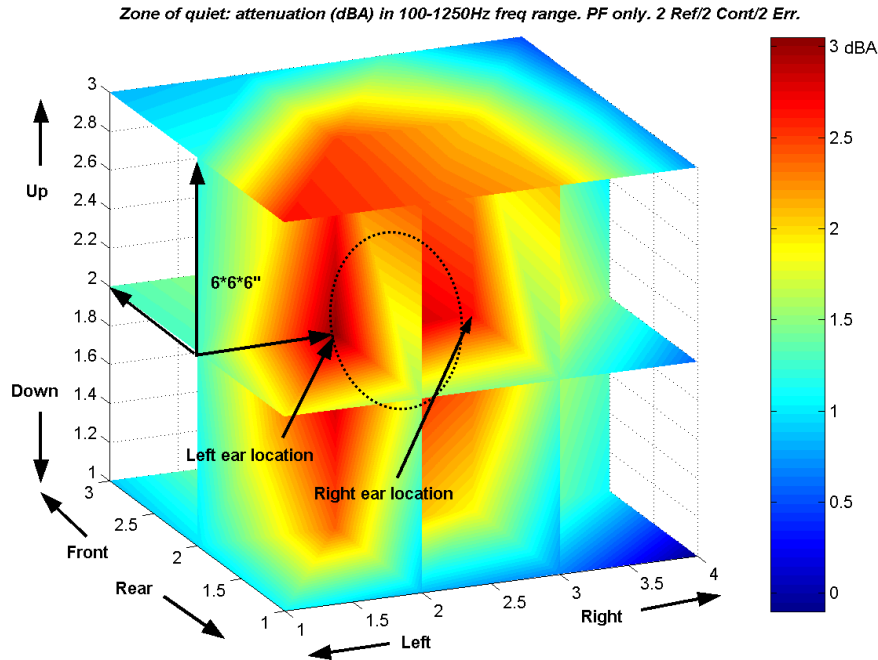


Figure 6.26 Spatial control performance. Processor fan noise.

We can finally analyze the control results when the power supply block fan, the processor fan, and eventually the hard drive (it does not affect the results) are simultaneously operating (c.f. Figure 6.27, and Figure 6.28). The zone of quiet, although of similar shape and level (7dBA maximum) to the power supply fan case (c.f. Figure 6.25), seems smaller. We believe that once again, the spectral content of the primary sound field is partly determining the extent of the spatial control performance.

In these experiments, the control sources are between 1 and 1.5 meters away from the primary noise sources. At low frequencies, both the primary and secondary noise sources are likely to present monopole like radiation characteristics and their separation is small compared to the acoustic wavelength. Thus, good global control should be achieved. Conversely, as the frequency increases, the radiation pattern of the primary and secondary noise sources are more and more complex and their separation distance becomes comparable to the acoustic wavelength. Thus, we expect a complex interference sound field with multiple zones of attenuation or amplification of the noise level.

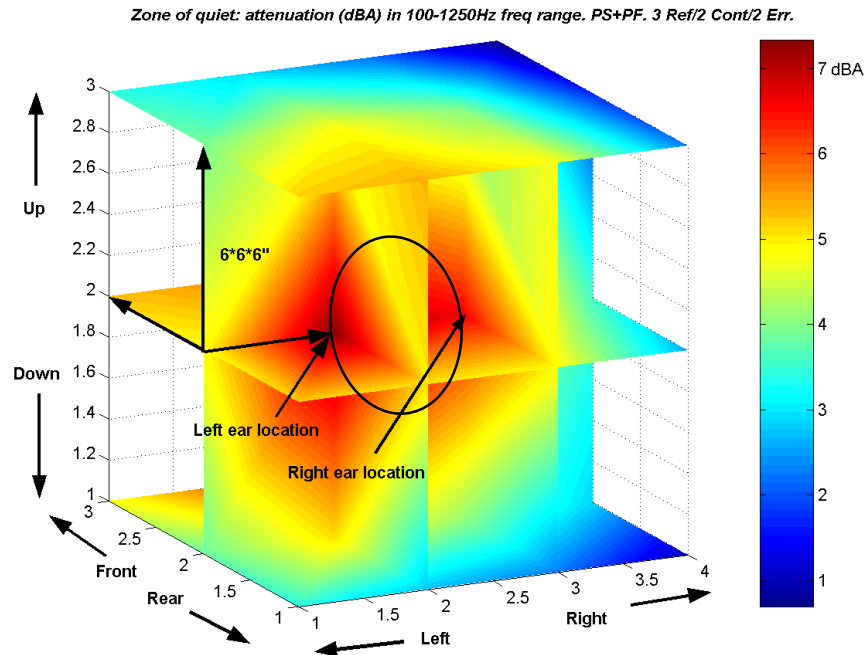


Figure 6.27 Spatial control performance. Power supply and processor fan noise.

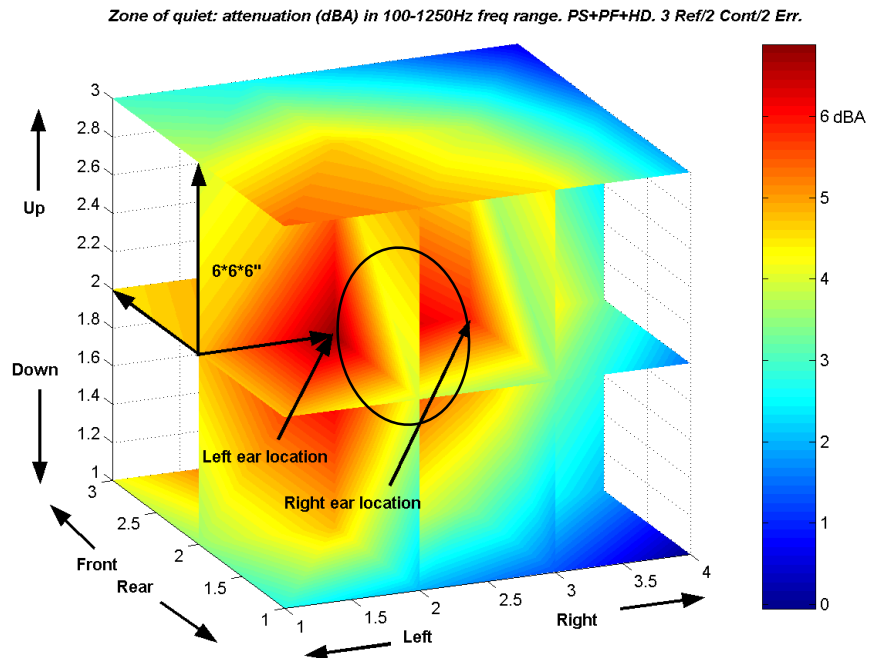


Figure 6.28 Spatial control performance. All sources operating.

Control performance at specific frequencies:

The conclusions from the previous section lead us to suspect that the spectral control of the disturbance to be controlled affects the spatial extent of the local ANC system. Our objective is now to further investigate this hypothesis by looking at the attenuation in narrow frequency bands. The following results are based on control experiments performed with all the primary sources (power supply block fan, processor fan, hard drive) simultaneously operating. The only difference with the results shown earlier is that the attenuation is computed in a narrow frequency band in order to observe the evolution of the size of the zone of quiet as the frequency increases.

We first have a look at the power spectrum at one of the error microphone locations. We chose the error sensor located at the typical position of the left ear of an operator. The results in Figure 6.29 correspond to the sound pressure level when the controller (three reference sensors, two speakers, two error microphones) is operating (solid red line) or not (dotted blue line). We observe that little broadband noise reduction is achieved, except in the 200-400Hz frequency range. On the other hand, many of the pure tones are well controlled, up to 1650Hz (upper frequency bound of control). Particularly, the power supply block fan BPF (at approximately 230Hz), which dominates the background noise level by 25dB, is almost perfectly attenuated at the user's left ear location. The second harmonic of this fan BPF (at about 460Hz) is not controlled as well, an observation that was already made in past experiments.

We then chose three narrow frequency bands to illustrate the frequency dependence of the size of the zone of quiet. The 3-dimensional plots shown in Figure 6.30 through Figure 6.32 represent the attenuation in the following frequency bands:

- Figure 6.30: 230-235Hz, power supply block fan BPF
- Figure 6.31: 905-920Hz, fourth harmonic of the power supply block fan BPF
- Figure 6.32: 1560-1570Hz, processor fan structure-borne noise radiation path with a pure in the upper-limit of the control bandwidth

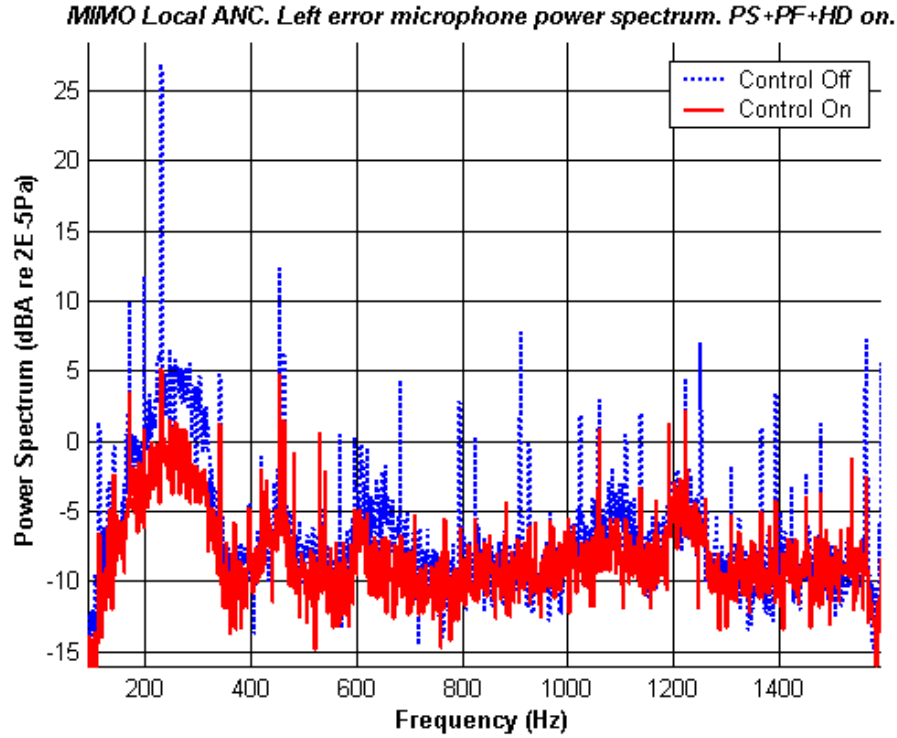


Figure 6.29 Local control performance (left ear error). All sources operating.

We first analyze the results from Figure 6.30. The maximum attenuation is very large since 18dB of reduction of the power supply fan BPF are achieved in the axis of the error microphones. Compared to the overall attenuation results (computed over a wide frequency range), the shape of the zone of quiet (surfaces with equal attenuation) is still oval, but its main axis is now in the axis of the error sensors. The size of the zone of quiet seems large compared to the overall results. At least 2dB of reduction are achieved at the monitor sensors the furthest away from the error sensors.

The acoustic wavelength at 250Hz is approximately 1.5 meters, which is about twice as much as the maximum spacing between any primary and secondary source. This could explain why the spatial extent of the zone of quiet is significantly large. We will observe the changes when the frequency increases.

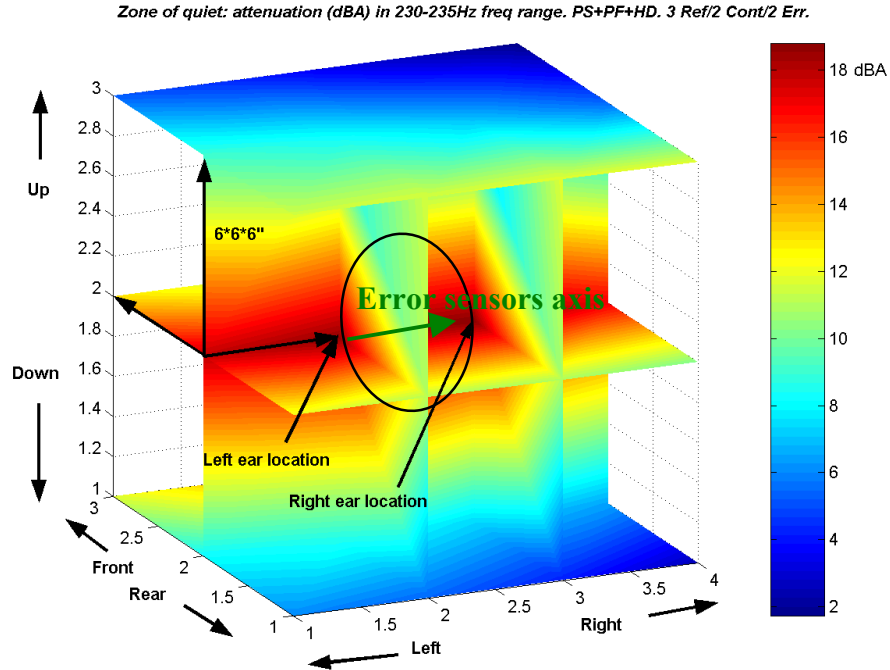


Figure 6.30 Attenuation of power supply fan BPF (230-235Hz).

We now analyze the results from Figure 6.31 where the attenuation is computed for the fourth harmonic of the power supply fan noise. Compared to the control performance at the power supply fan BPF, the results here are much degraded. For the first time, we observe constructive interference between the primary and secondary sound fields. Indeed, at some locations (blue color), there is an amplification of the sound pressure level when the controller is operating. Zones of constructive interference (cool color temperature) alternate with zone of destructive interference (hot color temperature). Thus, the size of the zone of quiet (defined as the spatial region around the error sensor where there is attenuation of the primary sound field) is smaller for the fourth harmonic of the power supply fan BPF compared to the results at the BPF.

The acoustic wavelength at 900Hz is approximately 35cm, which is about twice the spacing between the monitor sensors. It is thus reasonable that we observe mixed interference patterns since we “map” the sound field over at least a wavelength in all directions.

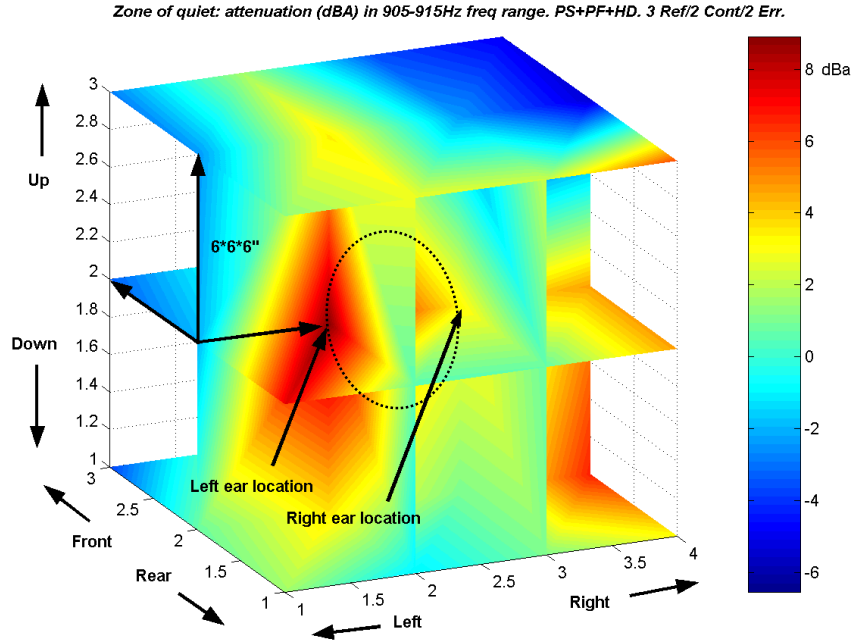


Figure 6.31 Attenuation of the fourth harmonic of the power supply fan BPF (905-920Hz).

We finally analyze the control results for a high frequency pure tone (at about 1560Hz) which is most likely related to the structure-borne radiation component of the processor fan noise. The results, shown in from Figure 6.32, are similar to what was observed at 900Hz (c.f. Figure 6.31). We clearly distinguish zones of constructive and destructive interference. Compared to the previous example, it seems that there higher the frequency, the more complex the interference pattern is. Since the delimitations in space between zones where the primary and secondary sounds fields are in phase or not is closely related to the acoustic wavelength, such phenomenon seems reasonable. The acoustic wavelength at 1500Hz is about 20 cm.

In this section, we have been able to identify the relation between acoustic wavelength and dimension of the zone of quiet surrounding the user's head. The next objective is study the effect of the number of secondary sources and error sensors on the dimension of the zone of quiet.

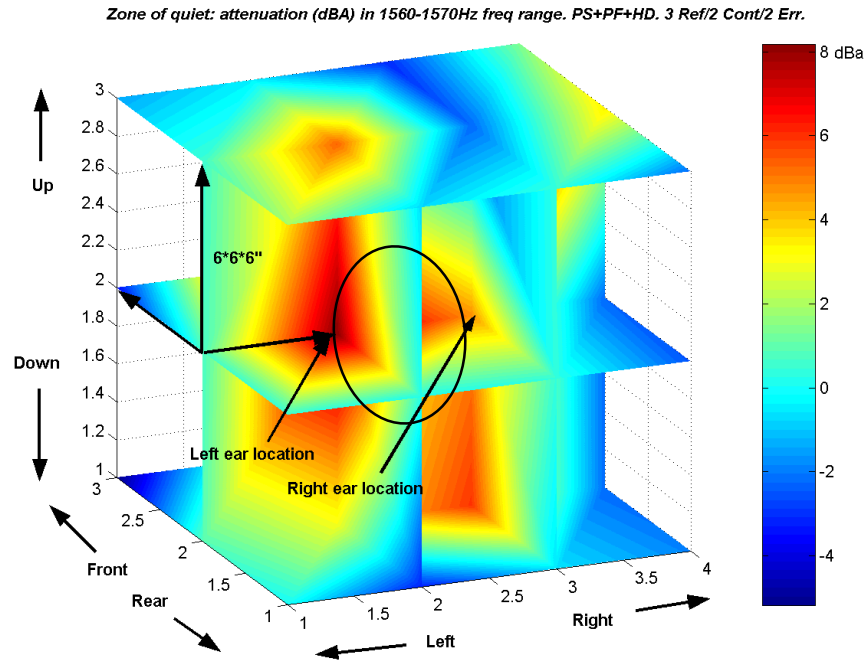


Figure 6.32 Attenuation of a high frequency pure tone (1560-1570Hz).

6.3.2.5. Effect of the Controller Complexity on the Performance.

a. Experimental Setup

For the following experiments, we leave the computer in the same simulated semi-reverberant environment as before. The computer is running under normal operating conditions with both fans and the hard drive powered up. Again, the PC is in its original configuration, except for the hard drive, which is structurally decoupled from the PC casing. Three reference sensors are used: one microphone next to the power supply block fan air outlet, one microphone next to the processor fan air outlet as well as one accelerometer attached to the PC casing (rear panel, near the fans). The controller settings all listed in Table 6.10. The three test configurations are listed in Table 6.11.

<i>Control system settings:</i>	<i>Value / Range:</i>
Reference signals analog filter:	Lowpass 3150Hz
Control signal analog filter:	Lowpass 1600Hz
Error signal analog filter:	Bandpass 20-1250Hz
Controller sample rate:	5000Hz
Digital identification filter:	120 taps
Digital control filter:	100 taps
Feedback removal filter	No
Error signal differentiation:	Yes
Coupling Control-Errors	100%

Table 6.10 MIMO controller settings for the control path complexity tests.

The controller is sampled at the fastest rate available in order to minimize causality problems (c.f. section 5.2.3.2). The parameters affecting the DSP load and thus the maximum sample rate allowed in the controller are the number of channels (reference, control and error), as well as the filters size.

The coupling scheme chosen assumes that each control actuator is coupled to each error sensor, which is true in the current experimental setup. Moreover, we assume that the impact of the secondary sources on the reference signals is minimal. Thus no feedback removal filter is implemented in order to save computational resources. Finally, to ensure a more stable control, the error signals are differentiated.

<i>Test case:</i>	<i>Number of secondary sources:</i>	<i>Number of error sensors:</i>
1	1 (Left loudspeaker)	1 (Left ear location)
2	2 (Left and Right loudspeakers)	1 (Left ear location)
3	2 (Left and Right loudspeakers)	2 (Left and Right ear locations)

Table 6.11 Test configurations for the control path tests in MIMO Local ANC.

b. Control Results

The results obtained for the three control configurations listed in Table 6.11 are shown in Figure 6.33 through Figure 6.35. We first look at the results when only the left side speaker is used to control the noise at the location of the user's left ear (c.f. Figure 6.33). Our first observation is that the zone of quiet is now centered on the unique error microphone.

The attenuation is thus very degraded from one ear location to the other with 7dBA achieved at the left ear vs. 2dBA at the right ear location. At monitor locations the furthest away from the unique error sensors, we also observe zones of constructive interference with an amplification of the primary source noise.

The control is also better in the direction of the control speaker (pointing toward the left ear). Once again, we might be observing a phenomenon due to the anechoic nature of the sound field. The noise seems to be better controlled in a conical region starting from the secondary source.

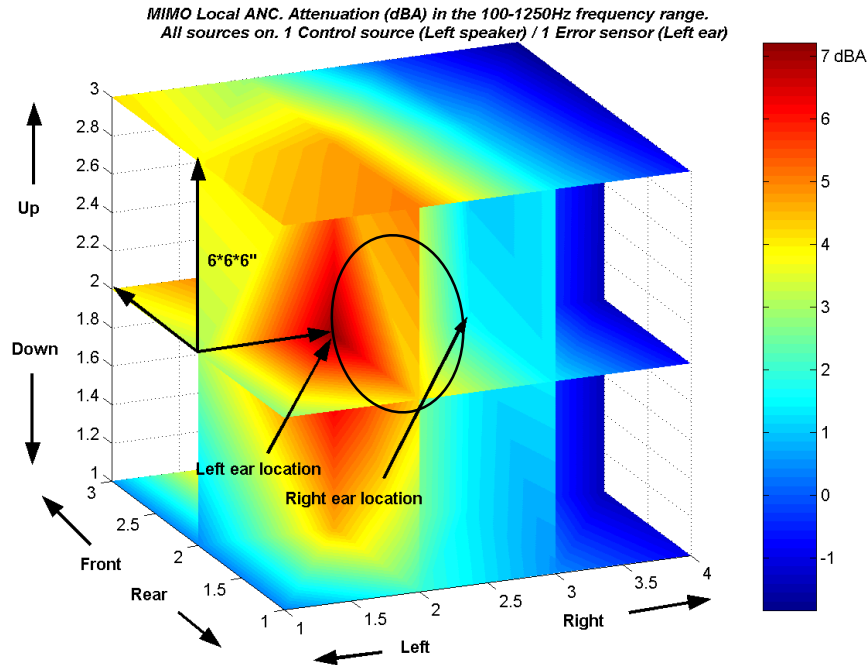


Figure 6.33 Spatial control performance with one speaker and one error.

The addition of a second actuator in order to control the noise level at a single error sensor location does improve the control performance as can be observed in Figure 6.34. Indeed, not only is the size of zone of quiet increased, but also the zones of constructive interferences are less important. However, we have to note that, theoretically, the use of more control sources than error sensors leads to an over-determined optimization problem with an infinite number of solutions [9]. Thus, the usefulness of such a control strategy is not obvious. We believe that the reason why the control performance is improved in this over-determined configuration is that the secondary noise field created by the two control sources better replicated the complexity of the primary sound field.

The control performance at the left and right ears still do not match (7 vs. 3 dBA). Thus, such control configuration is still not adequate since our objective is not only to maximize the size of the zone of quiet so that the user does not need to be confined in sweet spot, but also to achieve approximately the same level of control performance at both ears so that the control “feels natural”.

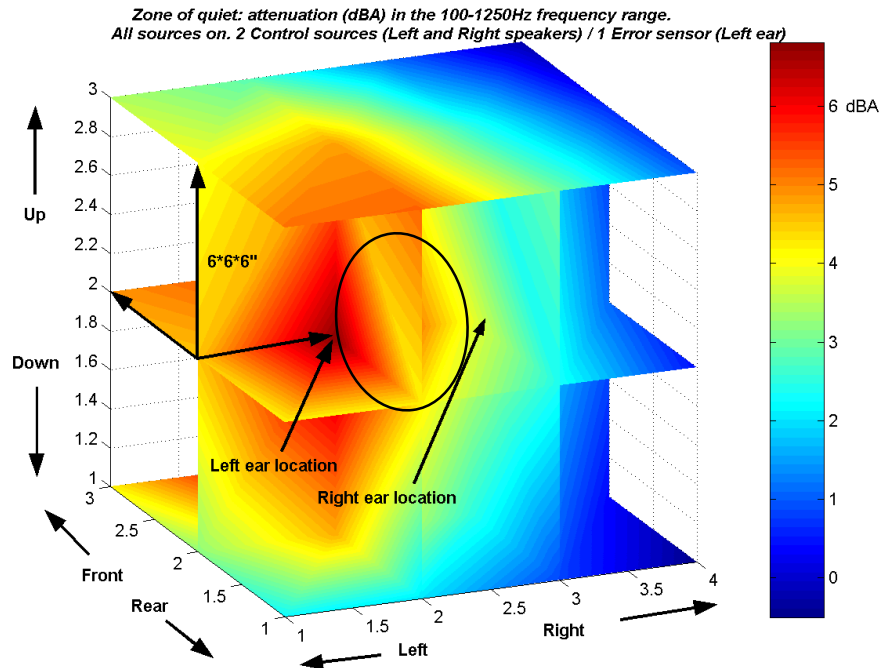


Figure 6.34 Spatial control performance with two speakers and one error.

One of the test configurations can satisfy both objectives listed above. It is the case of Figure 6.35 where two control actuators are used to minimize the sound pressure level at the both user's ear locations. The control performance at the two ears location is now similar with 6 to 7dBA of overall attenuation between 100 and 1250Hz. Moreover, the size of the zone of quiet is the largest obtained over such a broad frequency range of control.

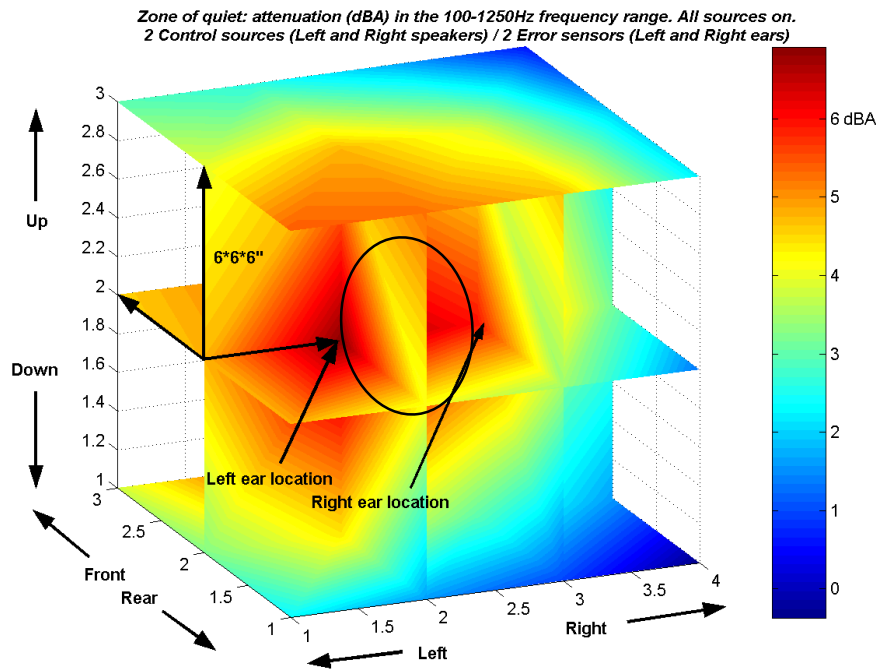


Figure 6.35 Spatial control performance with two speakers and two errors.

To conclude this section, we will observe that an effective method to increase the size of the zone of quiet surrounding the operator's head and to minimize the occurrence of constructive interference is to increase the number of control sources as well as error sensors. Due to time limitations, we could not investigate trade-offs between control complexity and control performance, nor optimize the multi-channel control system. However the current results are promising and further experiments should be carried out.

Chapter 7. Conclusions and Future Work

7.1. Conclusions

The purpose of this research was to find cost and size effective solutions to PC noise. Due to the important size constraints, it was shown that the implementation of passive noise control techniques is difficult. Consequently, a study of the potential of active noise control (ANC) to develop a cost effective, compact solution to the noise problems discussed above was initiated. It was conducted in four stages on a Dell Workstation computer donated by Intel Corporation. First, the stationary noise sources present in the PC were identified and characterized. Second, due to the wide frequency range of the primary disturbances and the multiple paths radiation, simple passive noise control techniques were investigated. This work would complement the performance of active noise control systems as well as facilitate their implementation. The third stage of this research was focused on the global control of fan noise using passive / active sound absorbers in ducts. Finally, local active noise control at the user head location using the on-board multimedia equipment was investigated.

The main conclusions of this research are as follows:

1. Three major stationary noise sources were identified: (i) the power supply block mounted in the PC casing with its integrated cooling fan, (ii) a chassis mounted fan dedicated to the cooling of the processor / heat sink unit, and (iii) a 7200RPM hard disk drive, that was part of the Dell Workstation 410 standard equipment. Non-stationary noise sources such as the floppy-drive, or CD-ROM drive were not investigated since they do not constitute targets for active noise control.

2. The fans and hard drive constitute the major stationary noise sources in the PC donated by Intel Corporation. The fans radiation consists of airflow induced broadband noise associated with pure tones. The pure tones are related to the fans Blade Passing Frequency. Both airborne and structure borne noise radiation paths are present due to the strong mechanical coupling between the vibrating noise sources and the casing. The broadband noise radiated is purely airborne, and present below 1000Hz. The pure tones are radiated via both airborne and structure borne noise radiation paths. The structure borne path is dominating for both the fans and hard drive noise at frequencies above 1000Hz. Thus, the casing is an efficient radiator over the frequency range of interest. Where as the fans are to be controlled using a combination of passive (structure borne path) and active (airborne path) noise control techniques, the hard drive noise is to be controlled passively due to the high frequency range (between 1000 and 3500Hz) and the dominance of a structure borne radiation path. All the noise sources identified in the PC have a similar sound power level, between 40 and 42dBA, leading to an overall sound power level of less than 45dBA between 100 and 6500Hz.
3. Passive redesign of the PC casing lead to very good performance results such as the elimination of the structure borne noise radiation path via mechanical decoupling of the vibrating noise sources. Generally, the high frequency noise was fully controlled for all the noise sources. Any broadband noise below 1000Hz needs to be attenuated with active techniques. The implementation of folded ducts at the air inlet and outlet locations yielded a significant attenuation of the fan noise radiation through one of the airborne paths, mainly due to changes in the fans acoustical load, from free field radiation to waveguide sound propagation. Nevertheless, no treatment was investigated to improve the casing transmission loss and reduce leaks. Thus, even though the air inlet and outlet were efficiently controlled, the overall transmission loss of the PC casing was not significantly improved. Overall, the complete redesign of the PC casing lead to a reduction of PC sound power level by 5dBA between 100 and 6500Hz. All the noise radiated above 1000Hz was completely attenuated.

4. The global active control of fan noise in short ducts was successfully implemented. An adaptive feedforward digital controller based on the Filtered-X LMS algorithm, and implemented on a Texas Instruments TMS320C30 DSP board was used. Three major technical issues, which are coherence, causality and control authority, were investigated. Coherence between reference and error signal was maximized by the use of windscreens on the microphones and by placing the reference sensor in the near field of the primary noise source. Maximizing the controller sample rate, minimizing the delays in the analog filters, and maximizing the separation distance between the reference and error sensors ensured the causal operation of the controller. The use of a reference microphone located in the upstream section of the duct, in the acoustic near field of the primary source, lead to the best control of broadband noise propagating downstream the duct. It was also shown that pure tones were best controlled using an accelerometer mounted on the fan casing or a built-in tachometer signal as reference. When all the noise sources identified are operating simultaneously, broadband attenuations of 7 to 10dBA, between 100 and 1200Hz, were achieved at the error microphones located at the outlet of the 20 cm long ducts. Nevertheless, the implementation of a double input double output controller for the simultaneous control of inlet and outlet noise did not lead to a reduction of the overall sound power level radiated by the PC. This issue is under investigation at the time of writing of this thesis. It was postulated that the lack of global performance was due to two problems. First, the PC casing has a low transmission loss. Second, the sound field (primary and secondary source contribution) radiated upstream the ducts, toward the interior cavity of the PC, is not controlled and can thus re-radiate in the exterior environment by an airborne radiation path through the casing.

5. Local active noise control at the head location, under a simulated semi-reverberant user environment was successfully implemented. A multi-channel adaptive feedforward digital controller based on the Filtered-X LMS algorithm, and implemented on an assembly of four Texas Instruments TMS320C40 DSP boards, was used. Control configurations involved three reference sensors (two microphones in the near field of the fans plus one accelerometer on the PC casing), up to two control sources (speakers located on a desk), and up to two error microphones at the ears location. Attenuations of 6 to 9dBA were achieved at the error sensors location when all the noise sources are operating simultaneously. The zone of quiet generated by the control system was evaluated by monitoring the sound pressure at 36 locations surrounding the error sensors. The monitored zone was thus a cubic volume of dimensions 18 by 18 by 18" surrounding the user's head. The power supply block fan BPF (250Hz) was attenuated by 6 to 18 dBA in a zone of 12 by 12 by 18" around the head. Nevertheless, zones of constructive interference were observed at higher frequencies (1000Hz, 1500Hz). The size of the zone of quiet could be significantly enlarged by increasing the number of control sources and error sensors.
6. The overall conclusion is that the current work is very promising. Even though a passive redesign of the PC casing helped to control most of the PC noise above 1000Hz, it was shown that active control was necessary and efficient for the noise radiated below 1000Hz. Both local and global active noise control strategies were applied with success in a PC and a large part of the remaining work is one of optimization.

7.2. Future Work

Due to the large scope of this research and time limitations, many problems encountered during this initial period could not be fully investigated or solved. Further work needs to be carried out to address several issues listed below:

1. The PC casing needs further modifications since its transmission loss was not improved at the current redesign level. Good passive control of the structure-borne noise radiation paths for the different sources was achieved. Moreover, the combined passive/active control of the noise radiated through the air inlet and outlet of the PC was also successful. However, the attenuation of these radiation paths lead to the discovery of paths that have not yet been targeted. Those include leaks as well as airborne radiation path through the panels of the PC casing.
2. The power supply block fan was not structurally decoupled from the PC casing. It is expected that better global active control of fan noise could be achieved if this isolation work was performed.
3. The active control of fan noise in duct was focused on the control of broadband noise, which leads to causality constraints and thus requires the design of significantly long ducts (at least 20cm in the current configuration). Shorter ducts could be designed when targeting pure tones exclusively. Moreover, feedback active noise control techniques are suitable for the control of broadband noise sources, and should be investigated. Eventually, a cost effective analog feedback controller could replace the adaptive digital controller.

4. The current global control strategy focuses on the noise control at the duct outlet exclusively. Eventually, the noise level inside the PC casing might be increased when the global control system is operating due to the propagation of disturbance as well as control noises upstream the ducts. Since the casing transmission loss is low, the effective performance of the global ANC system was poor. Thus, the design of a duct controlling both upstream and downstream noise radiation should be investigated or the casing transmission loss should be improved.
5. The current local active noise control system requires the presence of error microphones at the location of the ears, which is not practical. Virtual sensing technique should be investigated so that the error microphones can be relocated on the monitor, for example, but still minimizing the level at the user head location.
6. The current local active noise control system was tested in a simulated semi-reverberant user environment. Controller performance and robustness should be investigated in a more reverberant or noisier environment. Moreover, an optimization work regarding the number and location of error sensors and control sources should be carried out in order to achieve the most cost effective control performance.
7. Finally, the active noise control systems were physically implemented on dedicated DSP boards. Eventually, the on-board PC equipment should be used in order to reduce the implementation cost of active devices. The actual adaptive controller could be implemented on the DPS units of a sound card. One other solution would be to have the main processor perform the calculations as a background task. Major issues of required computational resources as well as data transfer bandwidth need to be investigated as part of a feasibility study.

References

- [1] Evans, G. W. and Johnson, D., "Stress and Open-Office Noise, ", *Journal of Applied Psychology*, **85** (5), 779-783 (2000)
- [2] Rayleigh, W. S., "Theory of Sound, ", Dover, New-York, 1945.
- [3] Coanda, H., "Procede de protection contre les bruits, ", French Patent No 722274. Filed: Oct. 21, 1930. Patented: Dec. 29, 1931. Published: March 15, 1932.
- [4] Lueg, P. , "Process of Silencing Sound Oscillations, ", US Patent No 2,043,416. Filed: March 8, 1934. Patented: June 9, 1936. Priority (Germany): Jan. 27, 1933.
- [5] Olson, F. and May, E. G., "Electronic Sound Absorber, ", *J. Acoust. Soc. Am.*, **25** (6), 1130-1136 (1953)
- [6] Conover, B., "Fighting Noise with Noise, ", *Noise Control*, **94** (6), 61-63 (1956)
- [7] Kido, K., "Reduction of Noise by use of Additional Sound Sources, ", *Proceedings of Inter-Noise 75*, Sendai, Japan, 547-650 (1975)
- [8] Chaplin, G. B. B. and Smith, R. A., "Active Methods of Canceling Repetitive Vibrations, ", U.K. Patent 19717/76, 1976
- [9] Nelson, P.A. and Elliott, S.J., "Active Control of Sound", Academic Press, London, 1992.
- [10] Fuller C.R. and Von Flotow A.H, "Active Control of Sound and Vibration, ", *IEEE Control Systems Journal*, **15** (6), 9-19 (1995)
- [11] Burgess, J. C., "Active Adaptive Sound Control in a Duct: A Computer Simulation," *J. Acoust. Soc. Am.*, **70** (3), 715-726 (1981)
- [12] Widrow, B., Shur, D. and Sohaffer, S., "On Adaptive Inverse Control, ", *Proceedings of the 15th Asilomar Conference on Circuits, Systems and Computers.*, 185-195, Pacific Grove, CA (1981)
- [13] Bendat, J.S. and Piersol, A.G., "Random Data: analysis and measurement procedures." 2nd edition, Wiley, New York, 1986.
- [14] Bendat, J.S. and Piersol, A.G., "Engineering Applications of Correlation and Spectral Analysis." 2nd edition, Wiley, New York, 1993.
- [15] Oppenheim, Alan V., Willsky, Alan S. and Nawab, Hamid S., "Signals and Systems." 2nd edition. Prentice Hall, New Jersey, 1997.

- [16] ANSI S12.30-1990, "Precision Methods for the determination of Sound Power Levels of Broad-band Noise Sources in a Reverberant Room," Acoustical Society of America, New York.
- [17] ISO 3741:1988, "Acoustics—Determination of Sound Power Levels of Noise Sources—Broad-band Sources in Reverberation Rooms," International Standardization Organization, Geneva, Switzerland.
- [18] ANSI S12.35-1990, "American National Standard Precision Methods for the Determination of Sound Power Levels of Noise Sources in Anechoic and Hemi-anechoic Rooms," Acoustical Society of America, New York.
- [19] ISO 3745:1997, "Acoustics—Determination of Sound Power Levels of Noise Sources—Precision Methods for Anechoic and Semi-anechoic Rooms," International Standardization Organization, Geneva, Switzerland.
- [20] F.J.Fahy, "Sound Intensity" 2nd ed., E&FN Spon, London, 1995.
- [21] ANSI S12.12-1992, "Engineering Method for the Determination of Sound Power Levels of Noise Sources Using Sound Intensity," Acoustical Society of America, New York.
- [22] ANSI S1-12-1994, "Instruments for the Measurement of Sound Intensity," Acoustical Society of America, New York.
- [23] Fuller, C.R., Elliott, S.J and Nelson, P.A., "Active Control of Vibration", Academic Press, London, 1996.
- [24] Sen, M.Kuo. and Morgan, R.Dennis., "Active Noise Control Systems: Algorithms and DSP Implementations", John Wiley and Sons, New York, 1996.
- [25] Beranek, Leo L. and István L. Vér., "Noise and Vibration Control Engineering: Principles and Applications", John Wiley and Sons, New York, 1992.
- [26] Kinsler, Lawrence E., Frey, Austin R., Coppens, Alan B., Sanders, James V. "Fundamentals of Acoustics", 3rd edition, John Wiley and Sons, New York, 1982.
- [27] Fahy, Frank J., "Foundations of Engineering Acoustics", Academic Press, 2000
- [28] Dickason, Vance, "Loudspeaker Design Cookbook", Audio Amateur Pub, 2000
- [29] Brillouin, L. and Sommerfeld, A. "Wave Propagation and Group Velocity", Academic Press, 1960

Vitae

Arnaud Charpentier was born on December 18, 1975 in Romilly-sur-Seine, France. After graduating from high school in 1994, he enrolled the University of Technology of Compiègne, France. After two years of general engineering studies, he specialized in Acoustic and Vibrations and earned his Mechanical Engineering degree in 1999. In 1998, Arnaud joined the Vibration and Acoustics Laboratories at Virginia Tech, where he started a master degree in Mechanical Engineering, under the advisory of Dr. Chris R. Fuller. His research project was dealing with the Active Control of Noise Radiated from Personal Computers. Arnaud joined the company Vibro-Acoustic Sciences in 2000 as an applications engineer. He has been involved in consulting work with the automotive industry regarding noise and vibration prediction problems involving experimental as well as analytical Statistical Energy Analysis. His current research projects deal with the use of statistical methods for the optimal design of active noise control systems or auralization.

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