THE ACTIVE CONTROL OF ACOUSTIC NOISE IN A SMALL ENCLOSURE

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A THESIS SUBMITTED TO

THE UNIVERSITY OF LONDON

FOR THE DEGREE OF DOCTOR OF PHILOSOPHY

PHYSICS DEPARTMENT
ROYAL HOLLOWAY AND BEDFORD NEW COLLEGE
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JUNE 1988

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This thesis describes the theory and implementation of a system for the active control of acoustic noise in a small enclosure ('small' infering that only a small number of acoustic modes dominate the reverberant field in the enclosure)

The theory for a multichannel active noise control system is developed. A system could consist of a number of detectors and cancellation sources controlling the field at a number of monitor positions. The simplest system comprises a single detector and a single source and is capable of controlling the field at one position. The controller for this system could consist of a pair of electronic filters, one between the detector and the source in parallel with another cancelling the feedback from the source to the detector. It is shown that for any configuration of transducers the required controllers can be realised by repeatedly using the same filter pair as described above.

A particular study is made of the active control of reverberant fields. A system was successfully implemented to partially control the reverberant field in a small

enclosure (0.5 x 0.6 x 0.7m). This consisted of a single detector microphone and a single control loudspeaker controlling the field at a single monitor microphone.

The system was controlled with two finite impulse response filters realised using a Texas Instruments TMS32020 microprocessor accessed via a Ferranti PC860XT personal computer. The first two modes of the reverberant field were successfully attenuated.

In summary, the theory for a multichannel controller has been developed and the simplist case tested. The importance of this is that the basic unit of a multichannel controller has been successfully implemented and this unit could be replicated as the basic building block for more complex controllers. This facilitates the implementation of controllers to attenuate the sound field at a number of points in a practical enclosure.

ACKNOWLEDGEMENTS

I am very grateful for the support and direction of Dr. Stuart Flockton who supervised this work.

I also thank Nick Abbott and Mike North who played a large part in the design and construction of the hardware used and also Tom Gurrie for our exchanges of information. All of the above made up the electroacoustics section of the Physics department of Royal Holloway and Bedford New College, London University.

This work was funded by the Science and Engineering Research Council under a CASE award with Plessey Research (Caswell) Ltd. Acknowledgement also to Tony Salloway, my industrial supervisor for his interest and encouragement.

CONTENTS

LIST OF FIGURES

LIST OF SYMBOLS

A list of the most common notation used in this thesis;

a	complex mode amplitude.
A_i	Mode amplitude of i^{th} mode.
ψ_{ij}	Characteristic function of the j^{th} mode at the i^{th} monitor position.
a b d	Dimensions of enclosure.
A	Matrix of transfer functions between noise source and monitor microphones.
B	Matrix of transfer functions between noise source and detector microphones.
C	Matrix of transfer functions between control speakers and monitor microphones.
E	Matrix of transfer functions between detector and monitor microphones.
Ex Ey Ez	Directional coupling coefficients of a mode.
F	Matrix of transfer functions between control sources and detector microphones.
k	Damping factor of a mode.
h_i	Filter coefficients.
L0 A0	Noise source loudspeaker and its amplifier.
L2 A2	Control loudspeaker and its amplifier.
M1 A1	Detector microphone and its amplifier.
M3 A3	Monitor microphone and its amplifier.
P_{xyzwt}	Sound pressure at a point.

T Matrix of transfer functions of the electronic controllers.

Q Volume velocity of a sound source.

$l\ m\ n$ mode orders in x, y and z directions.

r or ρ Density of air.

T_{60} Reverberation time.

t_n Decay time of a mode.

v Velocity of air.

V Volume of enclosure (m^3).

w Angular frequency.

w_n Modal angular frequency.

$x0, y0, z0$ Position of sound source.

x_1 Numerically generated swept sine signal.

x_2 Swept sine signal used as the excitation signal in the acoustical measurements.

y_{10} Signal response at the detector microphone due to noise source loudspeaker being excited by x_2.

y_{30} Signal response at the monitor microphone due to noise source loudspeaker being excited by x_2.

y_{12} Signal response at the detector microphone due to control loudspeaker being excited by x_2.

y_{32} Signal response at the monitor microphone due to control loudspeaker being excited by y_{10}.

yd_{12} Signal response at the exit from the digital feedback path due to being excited by x_2.

Z impedance.

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1 INTRODUCTION

Chapter one consists of a broad introduction to active sound control and describes the type of situations to which it may be applied. In particular, recent work on the active control of the reverberant field in small enclosures is referred to and the particular areas of interest of the subject are highlighted. These topics are covered within the framework of a review of the material presented in this thesis.

1.1 SITUATION

The active control of acoustic noise aims to attenuate unwanted low frequency sounds by using a number of loudspeakers to produce sound in antiphase to the unwanted sound. An active control system consists of a number of sensors and excitors, the former sensing the displacement, velocity or pressure etc. and the latter exciting the system to control the measured quantity. An active sound control system can use microphones to sense the sound pressure and loudspeakers to excite the sound field. The signal sensed by the microphones needs to be appropriately processed before being used to excite a control loudspeaker. A new generation of commercially affordable digital microprocessors with high processing speeds enable the signals to be suitably processed and the active systems to be controlled at costs that are sufficiently low to be attractive to industry.

Passive damping, such as acoustic tiles and partitions, operates by dissipating the acoustic energy by friction and is most successful at high frequencies. It becomes inefficient as the wavelength of the sound to be attenuated becomes much greater than the dimensions of the absorbing material. At higher frequencies an active system would be required to operate with high timing accuracy to give a pressure wave almost exactly out of phase to the unwanted sound, thereby greatly increasing the complexity of the system but an active sound control system that operates at low frequencies can act as an appropriate complement to passive control methods.

Ideally it is desirable to implement control systems which adapt to changes in the environment and would be capable of operating in many differant practical situations. Changes in the system to be controlled may be brought about by changes of temperature, movement of people or objects within the enclosure. It would also be an advantage to include some measure of the sensitivity of the human ear in the control system perhaps by relating the attenuation achieved to a phon scale. However, given these physical and psychoacoustic realities, it is first necessary to implement fixed (non-adaptive) control systems to gain an understanding of their requirements.

Much of the published work on active sound control systems has been concerned with the attenuation of the low frequency noise in ducts, particularly at frequencies below the cut off frequency of the first cross-mode of the duct. This is approximately a one-dimensional situation, thereby greatly simplifying the sound field to be controlled compared with a full three-dimensional system. Recently a commercial device has been marketed in the USA claiming to actively attenuate industrial fan noise (Ref. Nelson Industries Inc.). Another relatively simply situation is where a loudspeaker and microphone can be incorporated into a headphone making an active device designed to give attenuation only at the ear. This situation may, to a first approximation, be regarded as a zero dimensional case.

The sound field inside a room is much more complicated than these simple cases because of the multiple reflections and standing waves inside the room. The theory of how a three dimensional sound field may be actively attenuated is both less well understood and also is much more complicated to implement in practice.

1.2 COMPLICATION

Any sound field in an enclosure can be thought of as the sum of the propagating and reverberant fields. As previously mentioned, published study has concerned the control of propagating fields inside ducts and given sufficient processing power a one dimensional propagating field can be attenuated (Ref. Flockton). However, the control of the propagating field inside a room is much more complicated due to the multiple reflections involved and fundamentally requires that the control speakers be positioned in the line of the propagating field. Indeed, a propagating field is best controlled with a secondary source positioned as close as possible to the primary noise source. However, with multiple or large sources this may be difficult or impossible.

A practical situation can be envisaged of a machine operator requiring a quiet zone within a loud environment. A possible solution for this may be a general portable active sound control system consisting of a number of microphones and loudspeakers sampling and controlling the propagating field in a number of directions to produce a zone of attenuation.

In a practical situation it may be desirable to differentiate between the control of the direct and the reverberant sound fields. However, the principle of superposition can be applied to linear sound fields, therefore the propagating and linear sound fields can be considered seperately.

Although in reality the modes of a reverberant field will be complicated due to the shape of the enclosure, objects and people within the enclosure, resulting in shifts in the resonant frequencies and damping, the obvious simple situation to consider is a rectangular enclosure. A number of papers have been published concerning the active control of the low order modes of a reverberant sound field. Nelson (Ref.) has studied the problem in terms of what is the optimum attenuation that can theoretically be achieved with a number of loudspeakers controlling the field . The theory shows that substantial reductions in the net acoustic power radiated can be achieved if the control sources are within half a wavelength of the noise source. Bullmore (Ref.) has extended this theory to sound fields of low modal density by minimising the sum of the squared pressures at a number of sensor locations and has shown that attenuation close to the optimal levels can be achieved . Also, it is shown how attenuation can be achieved with control sources separated from the noise source by distances of greater than half a wavelength.

Many of the studies mentioned are theoretical and rely on prior knowledge of the noise source. Recent work (Ref. Elliott, Stothers and Nelson) demonstrates the practical attenuation of a harmonic sound field dominated by a few low order modes, the control system operating at a single frequency. Little material has been published concerning experiments on the active control of broadband noise within an enclosure.

1.3 SOLUTION

The theory and practical implementation described in this thesis demonstrates a number of aspects of active sound control applied in particular to the control of the low order modes inside a small enclosure. What follows is a breakdown of the structure of the thesis and the work presented.

Chapter two is divided into three sections, the first of which reviews the theory of the reverberant field inside an enclosure and shows how the dominant modes of the sound field need to be attenuated to produce worthwhile attenuation throughout the enclosure. Section 2.2 reviews the basic concepts of active sound control and indicates how the inherent feedback present in a system and the frequency responses of the environment and transducers may be compensated for. Section 2.3 combines and extends the known theory presented in the first two sections to show how an active system can operate to attenuate the low order modes of the sound field inside an enclosure. A general multichannel system is considered which can attenuate a number of modes and the requirements for the transducer positions are presented. An understanding of one way in which a control system can operate is presented by considering the attenuation of a single mode in the frequency and time domains.

Chapter three is also split into three sections; the first reviews the theory by which may be derived the required responses of the controllers for a general multichannel control system (first presented by Elliott and Nelson Ref.). The material is developed to give a new method whereby the digital controller for a multichannel control system can be readily realised and it is shown how the controller used to implement a single channel control system can be used as a basic unit, so that when used repeatedly a general multichannel control system can be implemented.

The first step in verifying the practical success of such a system is the practical implementation and testing of the basic unit of a multichannel controller. Therefore the next section describes the practical implementation of a single channel digital controller on a Texas Instruments TMS32020 microprocessor. The controller consisted of a pair of 128 point FIR filters; one in the forward direction positioned in the forward path from the detector microphone to the control speaker and the other in a feedback path to cancel the acoustic feedback.

Having developed a method of implementing a control system, section 3.3 presents the practical method whereby the digital control filters were derived for a single channel controller.

Chapter four presents the practical experiments carried out to verify the solution presented in section 2.3 and chapter 3. This particular approach shows how a single channel control system can be implemented to attenuate the low order modes of the sound field inside an enclosure. The experiments are also intended to support the theory that the single channel controller could be used as the basic building block of a multichannel controller and the first step in testing this theory is the testing of the basic unit.

Section 4.1 describes how a suitable test enclosure, practical apparatus and test conditions were configured to produce a situation in which a control system could be successful. The next section presents the measurements and analysis (theoretically presented in section 3.3) used to derive the digital control filters. Section 4.3 presents the results of measurements used to assess the degree of attenuation achieved by the control system and the results of the actual implementation of the system.

Section 4.4 presents the results of some additional measurements (i) Analysing the performance of the control system at higher sampling frequencies. (ii) Implementing a single channel control system configured to give attenuation at a different monitor position to the main experiment presented in the first three sections. (iii) A series of measurements is presented by which the digital controllers for a single detector, double source system can be derived. These measurements were carried out and the results of the implementation of the two channel control system discussed.

Chapter five discusses the main results and the conclusions that can be drawn from the work presented.

2 THE THEORY OF A REVERBERANT FIELD AND THE REQUIREMENTS OF AN ACTIVE SOUND CONTROL SYSTEM

This chapter consists of three sections. Section 1 is a review of the theory of the reverberant field inside a rectangular enclosure and section 2 consists of a review of the fundamental aspects of a simple active control system. In section 3 the material covered in the first two sections is brought together and expanded to gain a greater understanding of how the low order modes of a reverberant field may be actively controlled.

2.1 THEORY OF THE REVERBERANT FIELD INSIDE A RECTANGULAR ENCLOSURE

2.1.1 INTRODUCTION

This thesis is concerned with the active control of the low order modes of vibration of the reverberant sound field inside a rectangular enclosure. An appropriate introduction to the work presented is a theoretical study of the reverberant field inside a rectangular enclosure; in particular the sound field inside a rectangular enclosure generated by a single source within the enclosure. A practical implementation of an active sound control system operating inside such an enclosure is presented in chapter 4.

The modal nature of the reverberant sound field is discussed and the effect of the position of the sound source on the modal amplitudes is presented. These concepts are brought together in a single equation.

The sound pressure at a point inside a rectangular enclosure is expressed as the sum of pressure components from each of the modes of vibration of the air inside the enclosure. It is shown how each mode is excited to an extent depending on the source position.

The form of the amplitude and phase responses of the sound field inside the enclosure are explained. Finally, the effect on the sound pressure level of removing the dominant mode at any given frequency is shown and also the way in which the modes need to be controlled to attenuate the reverberant sound field in an enclosure.

2.1.2 THE THEORETICAL FORM OF THE REVERBERANT SOUND FIELD INSIDE A RECTANGULAR ENCLOSURE

A general equation for the pressure at a point in an enclosure due to a harmonic source is given by (Ref. Bullmore);

$$P\,(x,w) = \sum_{n=1}^{N} \psi_n(x)\, a_n(w)$$

where the excitation source and the pressure have a time dependence e^{jwt}. ψ are the characteristic functions of the enclosure and a are the complex mode amplitudes.

The reverberant sound field inside an enclosure can be thought of as being made up of the sum of the modes of vibration of the air inside the enclosure. Therefore the pressure at a point inside an enclosure consists of components from each mode.

The equation expressing the pressure at a point inside a rectangular enclosure can be found in many standard texts (such as Cremer or Pierce). The presentation that follows is intended to separate the relevant concepts within the equation.

The pressure at a point inside a three dimensional rectangular enclosure is given by the equation in figure 2.1.1. It is expressed as the sum of pressure components from each mode of vibration of the air inside the enclosue.

The pressure of each mode at a point is determined by three distinct factors. Firstly, the complex mode function; the mode amplitude and phase response to a given excitation. This is determined by the excitation frequency, the natural (modal) frequency and the degree of damping of the mode. This part of the equation shows that each mode acts as a second order system; (the response of an individual mode is studied in section 2.3).

Secondly, the complex source strength distribution. The position and driving function of the sound sources determines the extent to which each complex mode is excited. An obvious concept, sometimes overlooked in the mathematical theory, is that in a damped enclosure, unless an excitation source is present, a sound field will not exist.

This part of the equation relates the pressure generated to the volume velocity of the source. The source strength is always expressed by the element Q, the volume velocity of the source. The position of the source (the cosine terms) determines the extent to which it produces a pressure. This fits in with the fact that a baffled loudspeaker acts as a monopole source of strength determined by the rate of mass outflow from the source (rQ). Strictly the overall equation is valid for a point source. For generality, a summation is included to represent a number of sources; each source exciting the system at the specific angular frequency.

The complex mode function and the complex source strength distribution combine to determine the complex mode amplitudes (the magnitude and phase response of each mode).

The final part of the equation is the characteristic function of the mode. This takes into account the fact that the pressure at a particular point inherently depends on the position of that point. The function describes how the pressure amplitude of the mode varies with spatial coordinates.

The equation gives a greater understanding of the pattern of the sound field inside the enclosure; how the sound field is made up of the individual modes (standing waves) in the enclosure and what is the state of the sound pressure at a particular point.

The equation theoretically represents the sound field inside a rectangular enclosure where a sound source is exciting the field. It shows how the sources excite the modes of vibration and sets up standing wave patterns inside the enclosure (determined by the source positions and the complex mode shapes). These standing waves sum to give the pressure levels at all points in the enclosure.

2.1.3 THE MODAL NATURE OF THE SOUND FIELD INSIDE AN ENCLOSURE

The column of air inside a closed pipe can resonate around certain frequencies. When the air is excited a standing wave pattern is set up along the length of the pipe. Under ideal conditions when both terminations have identical impedances (different from the characteristic impedances of the pipe) the air in the pipe resonates at those frequencies where an integral number of half wavelengths fit into its length.

Consider a three dimensional enclosure: standing waves can exist in three mutually perpendicular directions; along the length, breadth or height of the enclosure. These waves are one dimensional, similar to the standing waves inside a pipe. The lowest frequency at which the air inside the enclosure will resonate is at the frequency for which one half-wavelength is equal to the longest dimension of the enclosure. Theoretically, the pressures associated with this standing wave across the breadth and vertical dimension of the enclosure are constant.

Standing waves can also be set up in two or three dimensions inside the enclosure; for example the pressure field varies across two dimensions as shown in figure 2.1.2. A standing wave can be present only when a source is present. However a standing wave is not necessarily equal to a mode; a mode is a mathematical solution to the wave equation with the appropriate boundary conditions, independent of a source being present. With a source present a sound field can be described in terms of different modal amplitudes and shapes.

A specific notation is used to distinguish the modes of the sound field. The first longitudinal mode referred to above is called the 1,0,0 mode. The 0,1,0 mode refers to the first mode in the y direction and the 1,0,2 mode has non-zero components in the x and z directions. The numbers specified in this notation are refered to as the order of the mode in that dimension. The 1,3,0 mode has order 1 in the x direction and order 3 in the y direction. The 2,0,0 mode is the second order mode (the second harmonic) in the x direction.

A mode is significantly excited (resonates) only when the excitation frequency is in the region of the natural frequency of the mode (the modal frequency) or the source distribution matches the mode shape well.

Exciting an enclosed volume of air at a particular frequency with a small source will excite all the standing waves within that volume. However, only those modes with natural frequencies in the region of the excitation frequency will be excited to any great extent and therefore these modes will dominate the sound field.

2.1.4 THE EFFECT OF THE SOUND SOURCE POSITION ON THE MODE AMPLITUDE

The amplitude of a mode may be defined as the amplitude of the maximum pressure fluctuations of that mode within the space being studied. It is a useful quantity to define the strength of a mode. However, we are concerned with attenuating the modal pressures; whether this is achieved or not can be determined by knowing the pressure at a single point in the standing wave.

The position of a sound source determines the extent to which that source can excite a particular mode. This can be illustrated by considering a simple system of a standing wave in a closed pipe (figure 2.1.3). For the first mode of the pipe the pressure at a distance x down the pipe is given by;

$$P = A \cos \frac{2\pi x}{X} \; e^{j\omega t}$$

The rate of change of momentum equals the force. In an acoustic pressure wave this determines that;

$$r \frac{dv}{dt} = - \frac{dP}{dx}$$

where v is the velocity of the air at a value of x and r is the density of air. Differentiating the pressure equation gives;

$$r \frac{dv}{dt} = \frac{2\pi A}{X} \sin \frac{2\pi x}{X} e^{jwt}$$

integrating this gives;

$$v = -j \frac{2\pi A}{wrX} \sin \frac{2\pi x}{X} e^{jwt}$$

The impedance of the standing wave in the pipe at a distance x is given by;

$$Z = \frac{P}{v} = j \frac{wrX}{2} \cot \frac{2\pi x}{X}$$

It can be seen from this equation that the impedance of the pipe is theoretically infinite at the ends where there is a wall and zero in the middle.

Towards the middle of the pipe the acoustic pressure variation gets less and less. In an ideal undamped standing wave the pressure does not vary at the exact centre of the pipe; a pressure node is said to occur here. A very small impedance is presented to a loudspeaker positioned on a node. The speaker is able to move easily for a small driving force. It can displace a relatively large volume of air without producing much pressure thereby operating primarily as a volume velocity source and not as a pressure source.

A loudspeaker at the ends of the pipe operates on a very large impedance; this impedance enables the speaker to get a hold of the air it is operating on thereby producing a pressure wave in the air; the speaker operates primarily as a pressure source. More specifically, whether the source acts as a velocity or pressure source depends on the ratio of the impedance of the source to the radiation impedance presented to the source.

Each frequency component of a sound source will excite a mode to the extent that the source couples into that mode. Likewise, a mode can only be detected to the extent that the detector couples into that mode. The equation of figure 2.1.1 incorporates these two concepts by expressing the pressure at a point in a rectangular enclosure in relation to the position of the sound source.

2.1.5 CALCULATION OF THE REVERBERANT SOUND FIELD INSIDE A RECTANGULAR ENCLOSURE

Having presented the equation determining the modal form of the sound field and the pressure at a point in an enclosure this equation can be used to determine the amplitude, phase and time response of the sound field inside the enclosure.

The sound field inside a rectangular enclosure of dimensions 0.5 x 0.6 x 0.7 m was calculated, using the equation in figure 2.1.1 for a position one fifth along a diagonal from the corner containing a point source (figure 2.1.4). Using an experimentally estimated damping the sound pressure at each frequency was determined by summing the pressure components from all modes up to order six. The modulus and phase pressure response were both calculated.

The degree of damping present in the enclosure determines the amplitude response of the modes of vibration of the sound field. The smaller the damping, the more prominent the reverberant field, the longer the reverberation time (T_{60}) of the enclosure and the peakier the amplitude spectrum of the response. A form of the pressure equation (Ref. Cremer) represents the damping component of the complex mode function in terms of the decay time of a mode (t_n) such that;

$$\frac{2w}{t_n} = 2 \text{ wn kn}$$

The decay time of a mode can be represented in terms of the reverberation time of the mode;

$$T_{60} = 6 \ln 10 \, t_n$$

The reverberation time of the enclosure used in the practical demonstration of chapter four was measured as being 1.6 s. Therefore the simulation was carried out using a decay time of 0.12 s for each mode (ie. the damping of all the modes was assumed to be the same).

The amplitude and phase response at the representative point are shown together in figure 2.1.6. This section continues by viewing the amplitude and phase response of the reverberant field in turn.

2.1.6 AMPLITUDE RESPONSE OF THE REVERBERANT FIELD AT A POINT IN THE ENCLOSURE

The amplitude response of the sound field shows the low order modes of vibration of the sound field (figure 2.1.5). The first peak represents the first natural resonance of the enclosure. Around this frequency the pressure field inside the enclosure is largely made up of a standing wave of half-wavelength equal to the length of the enclosure. The next peak represents the resonance for the height of the enclosure. At higher frequencies the modes become denser in frequency and overlap but at the lower frequencies the modes are distinct.

The figure shows the amplitude response due to a source of unit volume velocity. The linear amplitude response indicates the horrendous peaks that a control system has to deal with. The modal frequencies and the order of each mode with a natural frequency less than 700 Hz are listed in figure 2.1.6; these can be related to the peaks in the spectrum.

2.1.7 PHASE RESPONSE OF THE REVERBERANT FIELD AT A POINT IN THE ENCLOSURE

The complex part of the pressure equation determines the phase of the pressure response. The relative phase of the pressure in a standing wave changes by a half cycle (180 degrees) each half wavelength along a standing wave. A phase variation also exists due to the modal nature of the sound field; consider the phase response at the chosen point (figure 2.1.4). A plot of the amplitude response is shown in sequence with the phase response so the plots can be compared.

A more detailed explanation of the phase response is worthwhile. At resonance, the response of the system (the acoustic sound pressure level) is in phase with the excitation signal (in this case the velocity of the speaker cone); for an undamped system the frequency of maximum amplitude response and zero phase shift will coincide exactly. There is a rapid phase change on passing through the modal frequency; the less damped the system the more rapid the phase change. At frequencies above resonance the phase response begins to lag the excitation response more and more. However, at a frequency in between two modal frequencies the response is again in phase. There has had to be a reversal in the graph so that another phase lag could occur on passing through the next resonance. The pressure of this antiresonance can be seen by representing the pressure at a point as the sum of the pressures from just the two modes dominating the field at a frequency between adjacent modal frequencies. From the equation of figure 2.1.1;

$$P = \frac{A\ w}{w1\ k + j(w^2 - w1^2)} + \frac{B\ w}{w2\ k + j(w^2 - w2^2)}$$

A and B are constants relating to the complex mode amplitudes. This expression assumes the source was positioned at the origin so the cosine terms of the complex source strength distribution are constant for all modes. If the low order modes are considered then the characteristic function of a low order mode could have a similar magnitude for a general position inside the enclosure. For these reasons, to illustrate the shape of the phase response, the factors A and B can be considered equal and the pressure response considered proportional to;

$$\frac{1}{b + j(1-(w1/w)^2)} + \frac{1}{b + j(1-(w2/w)^2)}$$

where $b \ll 1$; the damping factors for each mode are taken as small constants (ie. it is a lightly damped system) of the same magnitude.

Ignoring the very small terms the imaginary part of this response can be represented by;

$$(\ 2 - (w/w1)^2 - (w/w2)^2\ )\ (\ 1 - (w/w1)^2\ )\ (\ 1 - (w/w2)^2\ )$$

There will be no phase lag when the imaginary part is zero; ie. when w = w1 or w2, the modal frequencies, or when;

$$w = w1\ w2\ (\ 2\ /\ (w1^2 + w2^2)\ )^{0.5}$$

it can be shown that this frequency lies between w1 and w2 verifying the existence of a frequency lying between the two resonant frequencies where the pressure is in phase with the source velocity. This has been shown for the phase response of a low order mode using a few assumptions; the phase response at higher frequencies will be more complicated but will have the same general properties.

In this numerical simulation the phase response was calculated at a point one fifth along a diagonal from the corner where the point source was placed. The pressure response at this point is not equal to the response at other points inside the enclosure. The acoustic pressure perturbation increases towards an antinode of a mode and the phase differs depending on the half cycle of the standing wave in which the pressure is monitored. If the phase response is monitored n half wavelengths from the source then the phase response (of an undamped reverberant field) will lag that of the velocity of the speaker cone by n half cycles.

In this simulation, for the low order modes below about 700 Hz the response has been calculated at a point in the same quarter wavelength as the source. The maximum phase difference is ninety degrees. However, at higher frequencies, the distance from the source to the detector can be greater than a quarter wavelength; more accurately, in the case of 2 and 3 dimensional standing waves, the source and detector lie on opposite sides of a nodal plane.

At higher frequencies the density of the modal frequencies increases (figure 2.1.4) and many modes dominate the field at a single frequency (Ref. Hough). At higher frequencies still, the field may be able to be regarded as a diffuse field with a uniform distribution of sound energy. This thesis is concerned with the active control of the low order modes of the reverberant field. It can be seen from the pressure equation how the modes can be regarded as independent systems. Therefore to attenuate the reverberant sound field it is necessary to attenuate each mode.

2.1.8 ATTENUATION ACHIEVABLE BY INDIVIDUALLY ATTENUATING THE LOW ORDER MODES

The principle of superposition applies to the reverberant sound field inside an enclosure; the sound field is composed of the linear sum of the modes of vibration of the enclosure. Because the modes are orthogonal and the principle of superposition applies then the reduction of the amplitude of one mode reduces the space averaged sound pressure level inside the enclosure. This quantity is representative of the overall sound level in an enclosure.

The effect of removing the locally dominant mode at all frequencies is shown in figure 2.1.7. The sound pressure levels are displayed on an arbitrary scale. It can be seen how the removal of a single mode leaves a pressure level which can be extrapolated from the response curves of the adjacent modes. At frequencies in between two modal frequencies the sound field is largely made up of contributions from the two nearest modes plus smaller contributions from all the other modes present in the simulation. If two modes are equally dominant at a particular frequency, then removing one of them will decrease the SPL by the order of 3dB as seen.

Removing the dominant mode at all frequencies demonstrates the significant reduction that can be achieved in the space averaged sound pressure level inside an enclosure. The noise heard by a listener would also be subjectively less annoying with the mode resonances removed.

Now consider the sound pressure level at one particular point, in this case one fifth distance along a diagonal from the source in a corner, and the effect of removing the most dominant mode at this point (figure 2.1.8). At frequencies in between two modal frequencies the two dominant modes vibrate out of phase; the mode of lower natural frequency lags the source and the mode with higher natural frequency leads. Therefore initially the modes are cancelling and removing one mode stops this cancellation and increases the SPL.

It is important to realise that this study has considered a reverberant field where the modes are distinct in frequency. Where the sound field has degenerate modes then removing one of a degenerate pair may not give such appreciable attenuation depending on the respective modal shapes (this can be seen at the intermodal frequencies). The situation is studied further in section 2.3.

2.1.9 SUMMARY

An active control system can only operate by working to attenuate the sound pressure level at a representative position or positions inside an enclosure. To attenuate the reverberant field inside an enclosure the system needs to be designed to control the field at a number of positions such that the dominant modes of the field are attenuated.

This section has presented the basic theory of the reverberant field inside a rectangular enclosure. The next section presents some basic concepts of how an active sound control system can operate, the information presented in these first two sections is then brought together in section 2.3 where the requirements of an active system to control the modes of vibration of the reverberant field are presented.

2.2 REQUIREMENTS OF AN ACTIVE SOUND CONTROLLER

2.2.1 INTRODUCTION

This section reviews some essential elements of an active noise control system ; it aims to present an understanding of the basic theory of active noise control. The theory is partly presented in the context of a system operating in a duct, which has been widely studied, since in a one dimensional situation the concepts involved can be grasped more easily.

The geometry of a basic active noise control system operating in a duct is presented and the essential concepts of digital signal processing are presented with particular reference to finite impulse response filters. Some basic control theory is covered to highlight the problems of instability. Finally the desired response of the controller is derived and control theory introduced at this stage to show how the problem of instability may be overcome by a suitable arrangement of FIR filters.

This information forms a solid introduction to the subject as a base for extending the ideas to active control in a small enclosure.

2.2.2 ESSENTIAL ELEMENTS AND ACTION OF AN ACTIVE NOISE CONTROL SYSTEM OPERATING IN A DUCT

Sound emanating from a noise source propagates through a medium in all directions in a manner depending on the characteristics of the source and environment. A noise source in a room or free space will send pressure waves in all directions but there may be a particular area where noise control is desirable. If we are concerned with only the direct field then the situation may be expressed as requiring the noise to be attenuated in a particular direction. The simplest version of this occurs in duct or pipe, an area which has received much study (Ref. Ross among others).

Publications on active noise contol in ducts have mostly been concerned with attenuating frequencies of half-wavelength greater than the largest cross sectional dimension of the duct; ie. those frequencies which ideally travel as plane waves down the duct.

A simple active noise controller in a duct could in principle operate by detecting the noise propagating down the duct and processing this signal so that, at a later time, a loudspeaker further down the duct can be caused to operate in such a way as to attenuate the noise.

The acoustic noise takes a time to travel from the detector to the loudspeaker, this time depending on the distance between the two. The responses of the transducers and the electronics connecting them in the control system itself must also cause the signal to be delayed. To achieve attenuation a signal must take as long to pass through the control path (and be inverted) as it takes the noise to travel along the duct. The controller cannot work for random noise if the acoustic delay is shorter than the unavoidable delay through the contol system.

The essential elements of the situation are shown in figure 2.2.1

1. A noise source (represented by a loudspeaker) causing sound waves to propagate down the duct.

2. A microphone detecting the sound to be attenuated. In this thesis this microphone is referred to as the detector; it is an essential part of the control system.

3. An electronic control system.

4. A loudspeaker driven in such a way as to attenuate the unwanted noise. This loudspeaker is referred to as the control speaker or secondary speaker.

5 A microphone monitoring the sound level, thereby monitoring the success of the control system. This microphone is referred to as the monitor. Its function is to indicate the sound pressure at a position and as such it is not an integral part of the contol system, as the system will operate without it. However it is a necessary element in establishing the way in which the control

system needs to operate.

Provided the right hand end of the duct is not perfectly reflective, then if the sound field at the monitor is identically zero, then the field will also be identically zero everywhere to the right of the monitor.

The idea that plane waves propagate in one direction only down a duct is a gross oversimplification that confused human active controllers for some time. The sound field in a duct (as well as the higher frequency cross modes) may predominantly consist of plane waves propagating down the duct from the noise source but waves travelling in the opposite direction will always be present from reflections at bends in the duct or at the open end (due to the change in impedance here). These reflected waves are significant and can be incorporated into the model as standing waves. The main study of this thesis concerns standing waves in enclosures and at this stage suffice it to say that it is important that transducers are not positioned at nodes of a standing wave if that frequency is to be controlled.

2.2.3 TRANSDUCERS

In a practical situation it is important that the transducers be sturdy enough to operate continuously within the environment and that their working frequency range encompasses the frequencies of the noise to be attenuated.

A microphone has a reduced response at low frequencies and if very low frequencies are to be controlled it is necessary to take particular care to choose a suitable microphone.

A loudspeaker, being an object with mass and stiffness, has a mechanical resonance; it approximates a mass-spring-damper system and a large phase change of the volume velocity (180 degrees if ideal) occurs on passing through the resonance frequency. The volume velocity of the loudspeaker diaphragm determines the strength of the equivalent monopole source. At frequencies below resonance the volume velocity of the speaker cone leads the driving voltage and above resonance it lags. The large phase change around resonance need not be a significant problem in the control system if it can be accommodated for by the digital controller. At frequencies near resonance the velocity of the cone is greatest for a constant input voltage. Therefore, for efficiency, it is desirable for the speaker used in a control system to operate near to the mechanical resonance of the speaker. The question arises as to over what frequency range the speaker can be said to be operating 'near' to resonance. Theory of a second order system indicates that below resonance the velocity of the speaker cone is proportional to frequency and above resonance it falls inversely to the frequency. The approximate form of this response is shown on a logarithmic scale in figure 2.2.2. It can be seen that if the speaker operates with a certain velocity at a frequency above resonance then the same velocity is also achieved at a frequency below resonance. Because of the logarithmic form of the response the lower frequency will be the same order of magnitude below resonance as the upper frequency is above. For example, if a speaker resonates at 200 Hz a suitable working range for the speaker to operate over may be from, say 20 to 2000 Hz. This frequency span can be defined as the range 'near' resonance over which the source will operate successfully.

The environment in which a source is to operate affects the output from the source; the sound power output in free space will be different from the output when the source is loaded by the impedance of some other system. The relevant concern in the implementation of active sound control systems is that the sound field to be cancelled may significantly influence the movement of the control source. It has been demonstrated theoretically (Ref. Silcox) "that the finite impedance of real sources in no way affects the linearity or the use of superposition"; ie. the loading effect on the source does not affect the sound field produced and the cancelling field can be regarded as being linearly added to the sound field to be cancelled. However, this thesis is concerned with the active control of the low order modes of vibration inside an enclosure; it is at these modal frequencies that the loading effect on the source will be greatest. The question as to whether or not in practice the environment loads the source to a significant extent remains unresolved, the implementation of a system indicating whether this is the case or not.

The transducers act as filters; they react differently to different frequencies. By far the most complicated filter in the system is the acoustic environment itself. As previously mentioned, standing waves are present in the duct; this constitutes frequencies being transmitted along the duct at varying amplitude and phase. This effect needs to be accounted for in the control system. To understand how this is achieved the effect needs to be represented mathematically. At this stage it is suitable to introduce some basic concepts used in digital signal processing.

2.2.4 DIGITAL SIGNAL PROCESSING USING FINITE IMPULSE RESPONSE FILTERS

The implementation of a suitable controller between the detector microphone and secondary speaker necessitates filtering the signal. The signal needs to be processed quickly (in real time); by the time any lumped data had been analysed the propagating sound would have left the region of interest. A generation of new, commercially affordable microprocessors has made a sampled data system the most attractive option for this task as easily alterable complex analogue filters are time consuming to design and construct. Digital filters have the advantage that;

1. They are easily implemented and changed.

2. They are not subject to drift or other physical problems; their performance is guaranteed.

3. The hardware system can be used for other purposes.

This thesis describes the implementation of an active control system using finite impulse response filters. Although other methods of filtering are available this method is the simplest to understand and simplist to use in practice.

A FIR digital filter works by sampling a signal at equal intervals of time and convolving the sampled time series with another time series (the digital filter tap weights).The resultant value of the convolution is output from the digital system at the next clock pulse. The analogue input signal is sampled by an analogue to digital converter (ADC); the action is mathematically equivalent to multiplying the continuous signal by a chain of delta pulses at intervals of T, the sampling period. The output signal y(i), sent to the digital to analogue converter

(DAC), is the result of convolving the sampled input signal x(i) with the sampled impulse response of the digital filter h(i).

$$y(i) = x(i) * h(i)$$

The process of convolving two time series does not directly provide information about the various frequency components in the time series. Standard texts (Ref. Newland) cover the Fourier transform; an operation transforming a time series into its frequency spectrum and vice versa. If the successive samples of a signal x(t) are a0, a1, a2 etc. then the signal can be represented in the Z plane as;

$$X(Z) = x_0 + x_1 Z^{-1} + x_2 Z^{-2} + \ldots\ldots x_i Z^{-i}$$

where $Z = e^{j\omega T}$ and T = the sampling period
The coefficient x_i is the sampled value of x at t = iT. Multiplication by Z represents a shift in time of +T. The convenience of this notation is that a simple relationship exists between the input spectrum X and the output spectrum Y.

$$Y(Z) = X(Z) \; H(Z)$$

This enables the transfer function H(Z) to be established more easily. This function, transformed to the frequency domain, represents the response of the system at a particular frequency.

If the time series are sampled at intervals of T seconds then the resulting frequency domain is defined at intervals of 1/T Hertz. The Nyquist criterion states that only frequencies up to a certain frequency can be resolved from sampled data. For an infinite length of data this Nyquist frequency is half the sampling frequency. Consequently it is necessary to sample at over twice the highest frequency of interest.

To prevent high frequency components being treated as lower frequencies (aliasing) the digital system needs to be band limited. A low pass filter (antialiasing filter) is needed at the entrance to the digital system. Another low pass filter (reconstruction filter) is needed at the exit from the digital system to remove the high frequency components which are present in the quantized signal.

The electronic controller needs to be designed to control the active system. At this stage it is instructive to review some basic concepts of control theory to understand the requirements of the controller.

2.2.5 CONTROL THEORY

Consider the action of the active control system in figure 2.2.1. Any sound generated by the control speaker will be detected by the detector microphone and the signal subsequently passed to the control speaker. This process is called feedback; the signal is continuously fed back to an earlier position in its path. In control theory terms a closed loop exists, a loop round which a signal can travel continuously. The howl sometimes heard in public address systems occurs because a feedback system has become unstable; i.e. the transfer function of the gain and phase round the loop are such that the signal level grows of its own accord, becomes very large and overdrives the speaker. This instability can occur at any frequency.

A simple model of a closed loop system is investigated to show how a closed loop system can become unstable. Consider the system shown in figure 2.2.3. The forward path of the system is shown to have a transfer function of A and the feedback path a transfer function of B. The signal at point x is given by;

$$x = X - B\,Y$$

The output signal Y is given by;

$$Y = A\,x$$

Therefore the transfer function of the complete system is given by;

$$H = Y/X = A/(1+AB)$$

This is the closed loop transfer function of the system, the frequency response measured across the closed loop.

For any system it is useful to have a meaure of how near a system is to being unstable; a system may be stable but be on the verge of being unstable. The closed loop transfer function does not give a measure of the degree of stability of a system. To quantify the degree of stability of a system a measure is needed of the transfer function AB.

The transfer function AB, known as the loop transfer function, is the response round the loop. To measure this response it is necessary to break the loop.

A system is unstable if the magnitude of the loop gain is more than one and the phase lag round the loop is an odd number of half cycles. The open loop response can be plotted in the complex plane (figure 2.2.3). It can be shown that the system will be unstable if, for increasing frequency, the response curve encircles the point -1,0. The margin of stability of a system is judged by how close the response curve approaches this point. Two criteria are used to determine the margin of stability. The gain margin is defined as the distance of the curve from -1,0 when the phase is an odd number of half cycles. The phase margin is defined as the angle between the negative real axis and the phase of the response when the gain is unity.

An important consideration for the acoustic system is that a time delay exists between the control speaker and the detector microphone. Whenever a significant time delay is present in the system then, because of reflections, there will be many frequencies that will be delayed by an odd number of half cycles (and at nearby frequencies the phase shift will be of the order of an odd number of cycles). Therefore the major criterion determining the stability of an acoustic noise control system is the amplitude gain round the loop.

In summary the digital controller needs to drive the speaker to attenuate the sound field and ensure that the system remain stable. To achieve this the digital controller needs to have a frequency response specific to the particular conditions it is operating in.

2.2.6 DESIRED CONTROLLER RESPONSES FOR AN ACTIVE NOISE CONTROL SYSTEM OPERATING IN A DUCT

Having developed the theory whereby frequency responses of a filter can be defined, then the desired response of the digital controller can be derived. The responses of various paths in the control system were labelled in figure 2.2.1;

E : The transfer function between the signal at the detector microphone and the signal at the monitor microphone; these signals are representative of the pressure levels at these positions. This path needs to be cancelled by a combination of paths T and C.

C : The transfer function between the signal at the

the control speaker and the signal at the monitor microphone.

T : The transfer function of the electronic controller.

F : The transfer function between the signal at the control speaker and the signal at the detector microphone. This feedback path causes a closed loop in the system which can be, and probably will be unstable at some frequencies. This acoustic feedback needs to be dealt with if the system is to work.

The required controller T is given by Ross (Ref.) among others and can be simply derived as follows;

The pressure at the detector microphone (P_1) is due to that resulting from the noise source directly (P_{o1}) plus the pressure due to the control speaker

$$P_1 = P_{o1} + T\,F\,P_1$$

Similarly, the pressure at the monitor microphone downstream is given by

$$P_2 = P_{o2} + T\,C\,P_1$$

Substituting in P1 from the first equation gives

$$P_2 = P_{o2} + \frac{T\,C\,P_{o1}}{1 - T\,F}$$

For zero pressure level at the monitor microphone then:

$$T = \frac{P_{O2}}{F\,P_{O2} - C\,P_{O1}}$$

The ratio of the pressure levels at the microphones due to the noise source alone (P_{O2}/P_{O1}) is the transfer function E. Therefore the desired response of the controller is given by:

$$T = \frac{E}{E\,F - C}$$

The filter required to cancel the unwanted sound and compensate for the feedback may be non-causal. A non-causal response may be thought of as needing to filter the signal ahead of time, ie. before it has reached the physical filter.

A understanding of the non-causality of a system may be seen by example. Consider the filter that would need to be implemented if the acoustic feedback did not exist. Refering to figure 2.2.1 the situation still exists of needing to filter the detected signal in such a way as to drive the control speaker to give cancellation downstream. It can be seen from the above equation that if the feedback F is set to zero then the response of the controller needs to be -E/C. The non-causality can be seen directly here; the system needs to model the transfer function of E and the inverse transfer function of C. C has a causal impulse response, of finite length, between the control source and the monitor microphone. However, the inverse impulse response needs to be such that, when convolved with the impulse response, the result is a pulse at a position t = 0 on the time axis. For this to be achieved, the time response of the inverse transfer function needs to extend into the negative section of the time domain and, depending on the complexity of the response, may need to be quite long. The inverse response implemented can only lie in the positive region of the time domain. This necessary shift in time needs to be available in the acoustic path. In other words the delay through the control system is comprised of the delay through the electronic components and a delay resulting from the digital filtering modelling the inverse impulse response of the control system. This total delay needs to be shorter than the acoustic delay for the system to cancel the direct sound.

However, the inverse response of any real system will be of infinite length thereby always ensuring that any implemented filter will be a causal approximation to a non-causal solution. this may be seen intuitively by considering the control system of figure 2.2.4. Consider a pulse leaving the control speaker; this pulse may, as intended give cancellation downstream. The pulse is also transmitted upstream and will be reflected from the upstream termination and will propagate downstream; this component of the pulse also needs to be cancelled by the controller. It is realistic that residual elements of this pulse will exist for a significant period of time thereby requiring the forward filter to be of substantial length.

It is important that the control system is stable; a system that would work well if it was stable, but is not, is of limited practical value. The understanding of how to implement successful systems will come from linking together the fields of digital signal processing and control theory. The approach taken here does this in a basic way.

The approach being taken at RHBNC (Ref. Flockton, Gurrie) has recently concentrated on splitting the controller into two paths; a forward path from the detector microphone to the control speaker in parallel with a feedback path (figure 2.2.4). This electronic feedback path is present to mimic and cancel the acoustic feedback path thereby stabilising the system and giving the engineer greater flexibility in the design of the forward path.

The response of the desired controller can be writen in the form;

$$T = \frac{1}{F - C/E}$$

Control theory states that a response of this form can be represented by incorporating a positive feedback path -F with a forward response of -E/C (figure 2.2.5). The digital path F intentionally cancels the acoustic feedback path. This thesis shows how a controller of this form was used to implement an active sound control system operating in a small enclosure.

2.2.7 SUMMARY

In section one the theory of the reverberant field inside a rectangular box was presented. Section two covered some basic concepts of an active noise control system and how such a system may be implemented. Section three uses and extends the material covered in the first two sections to show how an active control system may be used to attenuate the low order modes of the reverberant field inside a rectangular enclosure.

2.3 THE APPLICATION OF ACTIVE NOISE CONTROL TO ATTENUATE A REVERBERANT FIELD

2.3.1 INTRODUCTION

In the first two sections of this chapter the basic theory of the reverberant sound field inside a rectangular enclosure and the basic concepts of active noise control were presented. This section develops the theory of a reverberant field in greater detail and shows how an active system can attenuate the field. Although the material presented here may be deduced from theory, a number of different concepts are combined to show how a control system could operate and that the method can inherently work.

The requirements of the control system are presented along with the stipulations for the monitor, control speaker and detector positions. A control system consisting of a single detector and a single speaker controlling the field at a single monitor is considered. The form of the response of the controller is derived in the frequency domain and time domain and the intended operation of the control system is discussed. The practical implementation of such a control system is presented in chapter 4.

2.3.2 THE REQUIREMENTS OF AN ACTIVE CONTROL SYSTEM ATTENUATING A REVERBERANT FIELD

This sub-section develops the important requirements of how an active control system needs to attenuate the modes of the reverberant sound field inside an enclosure.

It is instructive to consider some different situations which will arise due to different noise sources. A single frequency noise source will drive all modes at that one frequency and the sound field will be dominated by those modes with modal frequencies nearest to the excitation frequency. Alternatively a broadband noise source will excite significantly all those modes which have modal frequencies within the bandwidth of the source. In all cases the reverberant sound field inside an enclosure will be dominated by a number of modes.

If the reverberant field is due to a single point source then the antisound can be produced using a single point source no matter how many modes are excited and substantial reductions can be achieved over all frequencies (ie. the direct field can also be attenuated) if the control source is within half a wavelength of the noise source (Ref. Bullmore). However, in general, a reverberant field will be generated from arbitrary noise sources and the modes will vibrate in an indeterminate manner.

In section 2.1 it was shown how the removal of the locally dominant mode at all frequencies can significantly attenuate the space average sound pressure in the region of that modal frequency. However, in that numerical simulation the modal frequencies were distinct in frequency; ie. the modes can be readily distinguished in the spectrum. At higher frequencies and in enclosures of different aspect ratio or shape some modal frequencies will be almost coincident and the mode responses will significantly overlap in the spectrum.

In practice it may be possible to control a number of modes with a single degree of freedom control system if the modes are sufficiently seperate in frequency. In general the modes of a reverberant sound field need to be controlled independently using a single control system to control each mode.

2.3.3 REQUIREMENTS OF THE MONITOR POSITIONS OF AN ACTIVE CONTROL SYSTEM ATTENUATING THE LOW ORDER MODES OF A REVERBERANT FIELD

Before considering how the mode amplitudes can be attenuated it is appropriate to consider how a control system can operate. Although it is required to remove the dominant modes of the sound field a control system cannot directly do this but can only operate to give attenuation at one or more monitor positions.

To attenuate the sound field over a volume the monitor positions need to be representative of the field over that volume. For example, an active system controlling the field at a downstream monitor results in the control of the plane waves travelling down the duct (section 2.2). The standing waves inside the duct will be attenuated if the transducers are positioned such that the waves can be detected, excited and monitored. However, a standing wave cannot be attenuated if a transducer lies on a node of the wave.

Likewise, in the case of a reverberant field, the monitors need to be positioned so that they see all the modes within the working range of the controller. This does not necessarily imply that the control system needs to resolve the amplitude of each mode but only that the monitors are positioned such that the signals from the monitors contain sufficient information that all the modes within the working range can be resolved.

The control system needs to isolate the modes in space or frequency. Furthermore the control system must be sure not to enhance other modes of the sound field which may not have been excited by the noise source. For example a mode may not be excited by the noise source if the source lies near a node, but if the modal amplitude lies within the bandwidth of the system then it must be taken into consideration. Therefore a general system controlling a reverberant field needs to be a broadband system and control all the modes within the working frequency range of the controller.

What follows is a mathematical investigation of the requirements of the monitor positions.

2.3.4 A MATHEMATICAL TREATMENT OF THE REQUIREMENTS OF THE MONITOR POSITIONS

Let the sound field in an enclosure be dominated by n modes and the amplitude of the i'th mode be $A_i(t)$. Let the pressure in the enclosure be sensed by n sensors and the pressure at the j'th sensor be $P_j(t)$. Then;

$$P_1(t) = \psi_{11}\, A_1(t) + \psi_{21}\, A_2(t) + \ldots\ldots + \psi_{n1}\, A_n(t)$$

$$P_2(t) = \psi_{12}\, A_1(t) + \psi_{22}\, A_2(t) + \ldots\ldots + \psi_{n2}\, A_n(t)$$

etc.

where ψ_{ij} is the characteristic function of the j'th mode at the i'th sensor position. It represents the fraction of the standing wave present at a position. $\psi = 0$ at a node and is maximum at an antinode.

The equations can be represented in matrix form;

$$P = \psi\, A$$

where the matrix ψ = characteristic function of the modes,
and the matrix A = mode amplitudes

The modal pressures at a point can be obtained from the inverse equation;

$$A = \psi^{-1}\; P$$

It can be seen how the characteristic function (or eigenfunction) of the enclosure needs to be known to determine the modal pressures. Methods are available to determine the characteristic mode shapes inside an enclosure (for example Ref. Kung) and in practice the modes shapes may need to be known in order to decide appropriate monitor positions. The mode shapes of the rectangular enclosure used in the practical demonstration of chapter 4 could be adequately determined by theory (section 2.1).

The important consideration in choosing the monitor positions is that the information present in the signal from the sensors is sufficient to adaquately define all the mode amplitudes within the working range of the control system. Each monitor needs to be placed in an independent position from the others such that the simultaneous equations presented above can be solved.

An example will illustrate the meaning of the term independent given above. In practice it will be desirable to monitor a mode at or near an antinode to maximise the pressure level detected and minimise the signal to noise ratio. However, consider the case of monitoring the 1,0,0 and 0,1,0 modes in a rectangular enclosure at positions given by the x,y,z coordinates 1,0,0 and 0,1,0 (antinode positions). It is shown in figure 2.3.1 how the matrix $\dagger$ becomes;

$$\begin{pmatrix} -1 & 1 \\ 1 & -1 \end{pmatrix}$$

The determinant of this matrix is zero, so the matrix cannot be inverted and the modal pressures cannot be resolved. This has occured because the chosen monitor positions were not independent: each position detected the same component of each mode. Mathematically, maximum independence is achieved by maximising the determinant of Ψ. A small determinant indicates that the simultaneous equations are very similar and in practice the noise present may make the solution unreliable.

This theory overcomes the problem of not being able to directly detect a modal pressure. If the characteristic functions of all modes within the bandwidth are included in the analysis and the pressure sensed at enough independent points then the modal amplitudes can be resolved.

It is possible to resolve n modes with n sensors provided they are appropriately positioned. However, using more sensors than the number of modes will give an excess of information to characterise the sound field; this may be desirable in a practical situation.

The matrix equation presented above contains information about the mode amplitudes; however it is not necessary that a control system determines the mode amplitudes in practice. Indeed, as previously mentioned the requirement is only that the information about the mode amplitudes be present in the signals.

2.3.5 REQUIREMENTS OF THE POSITIONS OF THE CONTROL SOURCES AND DETECTORS FOR AN ACTIVE CONTROL SYSTEM ATTENUATING THE LOW ORDER MODES OF A REVERBERANT FIELD

The reciprocal problem to detecting the modal pressures with an array of sensors is that of exciting modes in a given way with an array of speakers.

If a system response is dominated by a single mode then the response due to arbitrary sources will resemble that due to single point excitation because the principle of superposition applies. The modes are said to be orthogonal; altering the mode amplitude of one mode will not alter the amplitude of another. This implies that a single mode can theoretically be controlled by a single speaker. Likewise n modes can theoretically be controlled by n independent controllers.

If it is desired to control a number of modes which are distinct in frequency (section 2.1) the problem is theoretically easier; at frequencies near to a modal frequency the response is dominated by a single mode alone. The frequency response of a single speaker control system specified over the working range of the system may attenuate all the modal peaks in the spectrum (figure 2.1.7). This may theoretically occur because at and around the modal frequencies only one mode dominates the sound field.

The rectangular box used for the demonstration presented in chapter 4 had dimensions such that the low order modes were distinct in frequency. Even though the mode resonances are sharp, difficulties will be encountered if some modes are degenerate; ie. if two or more modal frequencies are very close.

The difficulty can be seen by considering the case of two modes of similar modal frequencies or the case of two modes both excited at a frequency in between their modal frequencies. The phase shift will be such that the pressure response of one mode (with the lowest modal frequency) will lead the velocity of the exciting signal and the other will lag. Such a situation indicates why the modes need to be controlled independently; a control signal at a particular frequency may attenuate one mode but necessarily enhance the other mode. Therefore it is necessary to control both modes simultaneously using two control sources.

Similarly, the number of microphones detecting the sound field needs to be equal to or greater than the number of degrees of freedom of the system. Each detector-speaker pair can be regarded as a single channel of the control system. In summary, the number of independent channels of a control system needs to be equal to or greater than the number of degrees of freedom of the sound field and each detector-loudspeaker pair needs to couple into at least one mode of the sound field. This is discussed further in section 3.1.

2.3.6 THE THEORY OF HOW A SINGLE MODE CAN BE ATTENUATED; FREQUENCY DOMAIN

Having developed the general requirements of the positions of the transducers in an active noise control system this section proceeds by looking at the action of the control system working to attenuate a single mode. To attenuate the reverberant field the control system needs to attenuate the modes of vibration of an enclosure; therefore it is appropriate to look at how a single mode could be attenuated.

Consider the control of a single mode; because of the large amplitude response of the mode it seems realistic that if a particular mode can be detected then there is a good chance of cancelling it. It is instructive to look at the cancellation needed in terms of the frequencies present; a typical amplitude response of a mode is shown in figure 2.3.2. To cancel the mode response it is necessary to superimpose a spectrum with the same amplitude response but in antiphase.

Consider the control system operating on a reverberant field in figure 2.3.3 analagous to the duct system of figure 2.2.1.

It can be seen how the magnitude response of the feedforward filter will be inversely proportional to the amplitude response of the feedback filter by analysing the control system with the transfer functions shown in figure 2.3.3. The feedback path is given by;

$$A1\ H1\ A2$$

The feedforward path is given by;

$$H2\ /\ H3\ A2\ H4\ A1$$

All paths through the acoustic system will have similar spectra, therefore it can be approximated that;

$$H = H1 = H2 = H3 = H4$$

Therefore, the feedback filter is given by;

$$F = A1\ A2\ H$$

and the feedforward filter is given by;

$$F = 1\ /\ A1\ A2\ H$$

Therefore the magnitudes of the responses are approximately the inverse of each other.

The detector microphone sees a signal of amplitude spectra shown in figure 2.3.2. The response of the path from the control speaker to the monitor will be of the same form; the antisound which reaches the monitor is the result of a signal passing twice through an acoustic system. The sound at the detector resulting from the noise source has passed through an acoustic system only once; therefore, to produce an antisound of the same spectrum the signal from the detector needs to pass through an electronic controller which models the inverse transfer function of a typical acoustic system. The amplitude response of the forward path of the electronic controller needs to model the reciprocal of the amplitude response of a mode (figure 2.3.4). The phase response required is such the the two signals are in antiphase at the detector.

2.3.7 THE THEORY OF HOW A SINGLE MODE CAN BE ATTENUATED; TIME DOMAIN

The theoretical response of the controller can also be determined in the time domain ie. the impulse response of the controller can be found. The impulse response of a system is the response of that system when it is excited by a unit impulse at time t=0; an impulse contains all frequencies and therefore the impulse response mathematically defines the complete response of a linear system. The impulse response of a lightly damped enclosure will be similar to that of the sum of a number of second order systems (the modes of the enclosure); it will consist of a number of superimposed damped sinusoids. A single decaying sinusoid and its Z transform is shown in figure 2.3.5. The inverse function is simply the reciprocal of the Z transform which directly gives the coefficients of the inverse filter. It can be seen that, theoretically the desired controller consists of only three coefficients.

In practice the theoretical impulse response of the control system cannot be implemented because the response is non-causal; one of the coefficients occurs in negative time on the time axis. Furthermore, inherent delays will be present in the control system; the delay of the loudspeakers and the electronics. These elements and the reality of the acoustic system will, of course, make the desired response of the digital filter more complicated than that derived above.

An active system operating on a propagating field or in a duct can sample the sound field and then attenuate the sound downstream thereby overcoming the inherent time delays in the system. This time may not be available for a system working to attenuate a reverberant field and the method whereby the modes can be attenuated needs to be understood.

A mode acts as a narrow band pass filter which responds well over a narrow frequency range. Therefore the time response of a mode to broadband excitation approximates a single sinusoid slightly varying in frequency over time. Consider the impulse response of a mode system (a second order system). The mode amplitude can vary between cycles by an amount determined by the damping in the system; the response of a lightly damped system will not vary much between cycles because little energy is dissipated in the system. the response of a heavily damped system will decay rapidly if no energy is added to maintain the oscillation.

Therefore the maximum rate of decay of a system is determined by the impulse response of that system. A practical example of a system whose response decays at this maximum rate could be a heavy press exciting the field in an enclosure with an impulsive sound. A system continuously excited by a broadband noise source will not decay at this rate. If significant attenuation can theoretically be achieved in the extreme case represented by the impulse response of the mode then the mode can be attenuated in all cases.

Because of the inherent delay in the system the cancellation signal will arrive after the noise signal. However, consider how a single impulse response could be actively attenuated; this is represented by the autocorrelation function of the mode. The first few cycles could not be cancelled because of the inherent time delay. However the controller is able to cancel the remaining cycles of the response (this is represented in figure 2.3.6); the impulse response of the controller is such that the response of the mode can be cancelled after a certain time. The reverberation present in the enclosure enables the control system to overcome the non-causality encountered if the walls were not present and the system were trying to attenuate the direct field; because the cancellation is occuring towards the tail of the impulse response the cancellation can be thought of as occuring in the far field. Alternatively, the cancellation can be viewed by considering how the control system would operate continuously; a particular cycle of a mode response can be cancelled by superimposing on it a processed version of a previous cycle.

The attenuation achievable will depend on the damping in the system; a heavily damped system will not ring on for long and sufficient time will not be available to achieve worthwhile attenuation. For a lightly damped system or when any system is being continuously excited then the amplitude and time period of the response will not vary much over time and cancelling a cycle with a processed version of a previous cycle will achieve worthwhile attenuation.

2.3.8 THE EFFECT OF DAMPING ON THE CONTROLLER RESPONSE

The amount of damping in a system determines the required length of the impulse response and the amplitude response of that system. The sound field in any real enclosure will be the sum of the direct and the reverberant fields. The relative importance of the reverberant field will be determined by the energy ratio of the reverberant field to the direct field. All physical systems are damped and damping implies the dissipation of energy; if energy is dissipated then energy must be continuously added to the system to maintain the vibration. The less damped the system, the less energy needs to be added to maintain the motion and the greater the amplitude response of the system. In an undamped system the pressure and velocity of the standing waves are a half cycle out of phase and their product, representing the energy transport is zero. The energy in the wave circulates with the harmonic motion, between potential and kinetic energy; if litte energy is dissipated then there can be little directional sound.

When designing a control system to attenuate a reverberant field it is instructive to obtain a measure of the length of the impulse response of the system. This will indicate the length of the impulse response of the digital controller needed. The control system that was implemented and is described in chapter 4 was realised with finite impulse response filters of 128 points. It is instructive to use this practical example to demonstrate how the length of a digital filter affects the degree of attenuation that can be acheived.

The sound pressure level of the impulse response of a reverberant field can be represented by the mean square pressure. The decay curve will be of the form shown in figure 2.3.7. The system described in chapter 4 was implemented at a sampling rate of 1 KHz and the group delay through the analogue elements of the control system was measured as being of the order of eight milliseconds. Therefore the control system can only attenuate the part of the decay curve beyond this time. The length of the FIR filters (128 points) determines that attenuation can only be achieved over this time length. These critical time intervals are shown in the diagram. The fraction of energy between these two intervals determines the attenuation achievable; this fraction is given by;

$$\int_{8}^{128} e^{-at} \Big/ \int_{0}^{\infty} e^{-at}$$

Therefore, from the decay curve of a reverberant field it is possible to determine the degree of attenuation that a system can achieve. For the enclosure used in the demonstration presented in chapter 4 the time taken for the curve to decay to a significantly small level was of the order of 300 mS (ie. 300 sample points at 1 KHz). Therefore, the fraction of energy that can be cancelled can be estimated by first deriving a value for the decay constant a. This can be found from the decay curve itself; by estimating a 60% decay to occur in 250 mS it is known that;

$$0.4 = e^{-0.25a}$$

$$a = 3.7 \ s^{-1}$$

Therefore

$$p^2 = A \ e^{-3.7t}$$

The fraction of energy that can be successfully cancelled is given by;

$$\sum_{8}^{128} e^{-3.7t} \ / \ \sum_{0}^{300} e^{-3.7t}$$

$$= 0.52$$

In order to attenuate a significant fraction of the reverberant field it is necessary to match the lengths of the digital filters to the impulse response of the acoustic system. It is described in section 4.1 how a small amount of damping was added to the test enclosure to shorten the impulse response of the acoustic system and increase the attenuation that could be achieved.

2.3.9 SUMMARY

This chapter has concluded by combining the theory of a reverberant field with the principles of active noise control to show theoretically how the low order modes of a reverberant field can be attenuated.

3. MULTICHANNEL CONTROLLERS; A METHOD TO REALISE THEM DIGITALLY AND A PRACTICAL IMPLEMENTATION OF A SINGLE CHANNEL DIGITAL CONTROLLER.

In section 2.3 it was shown how multichannel active control systems would be needed to attenuate a reverberant sound field. This is discussed in more depth in section 3.1 and the theory determining the desired frequency responses of a multichannel controller is presented. It is shown how the controller for a single channel system could consist of a pair of electronic filters; one between the detector microphone and the control speaker and the other cancelling the acoustic feedback from the loudspeaker to the detector.

The theory is extended to develop a method whereby the digital controller for a multichannel controller can be readily realised. It is shown how such a multichannel controller can be realised by repeatedly using the same filter pair as described above, thereby easing the practical apparatus needed to implement such a system.

The initial step in verifying this theory was the testing of the single channel digital controller which, it is shown, can be configured as the basic unit of a general multichannel controller. Section 2 therefore describes the practical implementation and testing of a digital controller on a Texas Instruments TMS32020 microprocessor. The controller consisted of two 128 fixed point FIR filters capable of operating a single channel active control system consisting of a single detector microphone and a single control speaker.

As well as suitable hardware and software to operate the control system it is necessary to derive the impulse response of the digital filters by some means. Section 3 presents the practical method used to derive the FIR filters. The method consists of a series of acoustical measurements on the control system and their subsequent analysis.

3.1 MULTICHANNEL CONTROLLERS

3.1.1 INTRODUCTION

Firstly this section reviews the theory which gives the required responses of the controllers for an active control system consisting of a number of detectors and speakers controlling the field at a number of monitor positions. A method is developed whereby a general active noise controller can be realised readily.

The material at the start of the section is largely a review of previously published work on single degree of freedom and multichannel control systems. This material is extended to develop what is, as far as is known, a previously unpublished method whereby the digital controller for a multichannel controller can be realised.

The practical need for multichannel control systems with a number of detectors and sources controlling the field at a number of positions is discussed. The response for a single source, single speaker system controlling the field at a number of positions is derived. This is developed to derive the response for a general multichannel controller. It is shown how the controllers for any multichannel control system can be realised by repeatedly using the same filter pair used in the single detector, single speaker control system.

3.1.2 THEORY OF A GENERAL MULTICHANNEL ACTIVE CONTROL SYSTEM

Many noise control situations can be simplified by controlling the noise at source. In active noise reduction it is advantageous to detect the noise field to be attenuated as near to the source as possible, more information about the sound field being available at this position and giving more time for the signal to be processed. Therefore study of active noise controllers has concentrated on systems which could detect the sound at some location and process the signal before activating the control spaekers to attenuate the field at other locations. The positioning of the transducers is important and the requirements for an active system operating in a small enclosure are discussed in section 2.3. However, given this, it is still necessary to have the correct controllers between the detectors and the sources to attenuate the field at the chosen poisitions.

An active system consisting of a number of detectors and sources controlling the field at a number of positions is shown in figure 3.1.1. A distinction is made between detector microphones and monitor microphones; the detector being the element which senses the sound field to be attenuated and the monitor being the position at which the system works to give attenuation.

It has been shown (Ref. Elliot and Nelson) how the responses of the controllers needed between the detectors and sources can be derived. Using the notation of figure 3.1.1 (and the single channel control system of section 2.2) where all terms are matrices of frequency responses (transfer functions) between the elements.

T : Desired controllers needed to give optimum attenuation at the monitors.

C : Responses between the control sources and the monitors.

F : Responses of the acoustic feedback paths between the control sources and the detectors.

E : Responses between the detectors and the monitors. $E = A/B$

A : Responses between the noise sources and the monitors.

B : Responses between the noise sources and the detectors .

Assuming the system is linear and using the principle of superposition, the signal at the monitor microphone is given by;

$$P_2 = A\ P_1 + B\ T\ C\ P_1$$

where, representing the noise source as a loudspeaker, the signal at input to the noise source speaker is given by P_1.

The sum of the squares of the outputs from the monitor microphones can be written as;

$$P_2^H P_2$$

where the superscript ^{H} denotes the conjugate of the transpose of the matrix. Applying the complex form of a standard minimisation formula the sum of the squares of the outputs of the monitors are minimised by a controller of response given by;

$$T = (C^H E F - C^H C)^{-1} C^H E$$

The notation of the individual terms of the matrices may be seen by example; the transfer function from the second source to the third monitor is given by c_{32} ie. the element in the third row and second column of the matrix C. The transfer function representing the response between the third source and the second detector, f_{23} is positioned in the second row and third column of the matrix F. Briefly, if the transfer function describes the response from the transducer i to the transducer j it is positioned in row j, column i of that matrix. The convention used here is that each column of a matrix refers to a different loudspeaker and each row refers to a different microphone.

Matrix E is slightly different; it represents the response between the detector and monitor microphones. This response cannot be measured without a noise source being present and each element is therefore represented by the response from the noise source to the monitor (A) divided by the frequency response from the the noise source to the detector (B).

Consider the case of each matrix having only one element, analogous to the duct system presented in section 2.2; one upstream noise source, one detector with a control source downstream from it and a downstream microphone to monitor the resultant sound passing down the duct. The matrix equation reduces to;

$$t = \frac{e}{e\ f - c}$$

At this stage it is worth emphasising that use of a single detector, single source control system is not necessarily confined to use in a duct. The main experiment presented in this thesis demonstrates the implementation and usefulness of such a system in a small enclosure.

It was shown in section 2.2 how this controller can be realised using a pair of fixed filters and thereby overcoming the inherent feedback problem. The controller can be represented as;

$$t = \frac{1}{f - c/e}$$

In this form it is easier to see the feedforward and feedback paths more readily. The electronic feedback path is given by f and necessarily needs to cancel the acoustic feedback path. The feedforward path is given by the reciprocal of the second term of the denominator, ie.

$-e/c$. This is a negative term to cancel, rather than reinforce the acoustic path. The feedforward and feedback components of the controller are perhaps obvious in this simple case. They are by no means obvious when designing a controller with a number of detectors, sources or monitors. Representing the desired controller in this form readily distinguishes the two components and this understanding of the simple case leads to an understanding of more complex situations.

3.1.3 THE NEED FOR MULTICHANNEL CONTROLLERS

The production of sound may be thought of as been brought about by (1) the initial disturbance, (2) the modification or amplification of that disturbance and (3) the radiation of the sound. These factors all need to be considered when planning the number and positioning of the detector microphones, control speakers and monitor microphones to be used in an active control system.

Only under certain circumstances will a noise source be definable as a point source; when operating in the far field of the source or in the reverberant field; in practical circumstances the noise sources and the subsequent modification of the sound by the enclosure may make the pattern of the sound field very complex.

In many cases it may be possible to obtain an independent estimate of the frequency of excitation, such as the case of a motor or fan operating at a particular frequency where the detection signal may be able to be tapped off from the mechanical system itself, thereby eliminating the acoustic feedback. Plane waves travelling down a duct or pipe may also be detected with a single microphone at a suitable position. However, where the sound field results from a source with many degrees of freedom, a large machine or resulting from a number of vibrating panels, then a number of microphones may be necessary to adequately define the field.

The major body of this work is concerned with active sound control in enclosures. It is instructive to consider the sound field in an enclosure as being generated by a number of loudspeakers. At the low frequency modes with which we are concerned here each speaker is small enough to lie within a quarter wavelength of a mode characteristic function. This model lets the speakers be considered as point sources with each independent speaker contributing one degree of freedom to the sound field. In the case of a direct field resulting from a single source, the pressure response at a single point has one degree of freedom. In the case of a diffuse field the number of microphones required to give attenuation over a given volume depends on the wavelengths of the sound to be attenuated and the volume over which the the attenuation needs to be achieved. However, the pressure at a point in a reverberant field can have many degrees of freedom depending on the number of independent sources. Therefore n points in a reverberant field can have n degrees of freedom.

A system with n degrees of freedom can only be adaquately defined by n or more independent outputs from that system, resulting in n simultaneous equations which can be solved. Under these conditions it is necessary to monitor the sound field with as many, or more independent monitors as there are degrees of freedom. In theory it is an advantage to overspecify the sound field by using more monitors than the number of degrees of freedom in the system, thereby being more sure of designing a system to attenuate the field over a given volume.

The number of control sources and microphones detecting the field in an active system are governed in a similar way to the number of monitors required. (see section 2.3.3) The number of independent control speakers and the number of detectors needed in an active system needs to be the same as or greater than the number of degrees of freedom of the sound field. In the case of the reverberant field the number of degrees of freedom is determined by the number of point sources in the enclosure. An additional requirement is that at least one microphone loudspeaker pair is coupling into each mode to be controlled.

In summary the number of channels of an active control system needs to be equal to or greater than the number of degrees of freedom of the sound field to be controlled. Each channel consists of a detector microphone and a control speaker.

The need for multichannel active controllers has been discussed. Given an active system consisting of a number of detector microphones, control speakers and monitor microphones the controllers connecting the transducers still need to be realised. At this stage some specific cases of multichannel controllers are presented to illustrate how the desired responses of the controllers can be obtained from the controller matrix equation.

3.1.4 AN ANALYSIS OF SOME SPECIFIC EXAMPLES OF MULTICHANNEL CONTROLLERS

Consider a controller consisting of one detector and one source controlling the field at two monitor positions. Let the matrices C and E have components c_1, c_2 and e_1, e_2. The required controller is then given by;

$$t = \left((c_1^*, c_2^*) \begin{pmatrix} e_1 \\ e_2 \end{pmatrix} f - (c_1^*, c_2^*) \begin{pmatrix} c_1 \\ c_2 \end{pmatrix} \right)^{-1} \qquad (c_1^*, c_2^*) \begin{pmatrix} e_1 \\ e_2 \end{pmatrix}$$

and hence

$$t = \frac{c_1^* \, e_1 + c_2^* \, e_2}{(c_1^* \, e_1 + c_2^* \, e_2) \, f - (c_1^* \, c_1 + c_2^* \, c_2)}$$

Extending this by induction to a controller with n monitors, the required response of the controller is given by;

$$t = \frac{\sum^{n} c_i^{*} e_i}{f \sum^{n} c_i^{*} e_i - \sum^{n} c_i^{*} c_i}$$

and hence

$$t = \frac{1}{f - \sum^{n} c_i^{*} c_i \,/\, \sum^{n} c_i^{*} e_i}$$

An analogy with the controller for the single monitor system shows that the feedback path needed in the controller has not changed. This follows because the acoustic feedback has not changed.

A simple but important statement: increasing the number of or moving the monitor microphones does not affect the acoustic feedback present in the active system. Designing the controller with independent feedback compensation infers that the monitors can be adjusted in position or number without altering the feedback compensation present in the controller. Only the forward electronic path needs to be changed. This suggests an advantage in designing stable control systems.

The feedforward path is the ratio of two terms. The first term is the sum, summed over the number of monitors, of the transfer functions between the detector and a monitor multiplied by the conjugate of the transfer function between the control source and the same monitor. This sum is divided by a second term. The second term is

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the sum of the transfer functions between the control source and a monitor multiplied by their respective conjugates. The feedforward path needs to have a response such that (denoting the forward paths as having frequency responses k);

$$e_j = \sum^{n} k_i \; c_{ij}$$

The complexity of the algebra of the desired controller responses increases faster than the number of elements in a control system; realising the more complicated controllers in an understandable format is not immediately obvious.

3.1.5 A METHOD TO REALISE THE CONTROLLERS FOR A MULTICHANNEL ACTIVE CONTROL SYSTEM

What follows is a development of a method whereby the controllers for a general multichannel controller can be readily realised by repeatedly using the same filter pair used in the single detector, single speaker control system. The theory presented so far in this section is largely a review and clarification of known results. The remainder of this section presents a method whereby the topology of the controller for any multichannel controller can be readily realised.

Consider the situation of wanting to control a two degree of freedom system; for example where it is required to control two modes of the reverberant field inside an enclosure and the resonance frequencies of the modes are near. It would be appropriate to use two speakers and two monitors giving two degrees of freedom in the active control system. The desired response of the two controllers is given by;

$$T = (C^H E F - C^H C)^{-1} C^H E$$

In this case the matrix C is square; it can be shown that if the matrix C is square then the equation factorises to give;

$$T = (C^H (E F - C))^{-1} C^H E$$

and using the associative rule of multiplication for matrices;

$$T = (E F - C)^{-1} E$$

Now, inserting the matrix elements and multiplying out the equation will result in the individual terms of the controllers. However, this is a lengthy process and increasingly more complex systems would require the use of computer algebra to solve the equations. The method also results in feedback terms which are more complicated than the acoustic feedback itself. This is a result of the structure of the control system model used up to this point; in this model a feedforward and a feedback path are used to model each individual controller (figure 3.1.3). It can be seen that in this case the acoustic feedback paths do not match the electronic feedback paths. Each electronic feedback path needs to mimic an acoustic feedback path plus an additional term resulting from the signal at the microphone passing through the other branch of the system and feeding back to the microphone via the other acoustic path. This phenomenon results in two simultaneous equations which need to be solved to determine the responses of the electronic paths. A simpler solution can be found.

It is instructive to review the simpler systems already presented and the way in which their controllers were realised, specifically the use of an electronic feedback path to cancel the acoustic feedback. We can extend this method to the two source controller. The acoustic feedback paths are from each control speaker (actually the point of exit from the digital system) to the detector microphone. The acoustic feedback paths add together at the detector microphone (actually the point of entry to the digital system). This feedback can be counteracted by modelling each acoustic feedback path electronically and summing the electronic feedback paths at an equivalent position to the acoustic feedback (figure 3.1.4) before the forward paths thereby cancelling the summed acoustic feedback.

The success of this method lies in the position where the feedback paths meet; ie. before the signal splits to enter the feedforward filters before each speaker. The advantage is that the electronic feedback filters have a simple response which only needs to model the acoustic feedback for that channel. These paths are causal ensuring that they can be adaquately modelled. This also eases the extraction of the desired response of the feedforward filters from the matrix equation.

With the feedback path cancelled the response of the forward path is given by;

$$K = - C^{-1} \quad E$$

This is a familiar expression; the forward response for the controller of a single detector, single source, single monitor system is given by - E/C. The expression above is simply the matrix form of the same equation. It can be expanded to give;

$$K = - C^{-1} \quad A \; B^{-1}$$

$$K = - \begin{pmatrix} c_{11} & c_{12} \\ c_{21} & c_{22} \end{pmatrix}^{-1} \begin{pmatrix} a_1 \\ a_2 \end{pmatrix} B^{-1}$$

$$K = \frac{1}{B\,(c_{11}\;c_{22} - c_{21}\;c_{12})} \begin{pmatrix} a_1\;c_{22} - a_2\;c_{12} \\ a_2\;c_{11} - a_1\;c_{21} \end{pmatrix}$$

The method is worth demonstrating for a more complex system. A system consisting of two detectors and two speakers is shown in figure 3.1.5. Again the electronic feedback paths are arranged to cancel the acoustic feedback and the matrix equation determining the controller reduces to the form given above.

It has been previously shown how a sound field with n degrees of freedom can theoretically be attenuated with an active system of n channels. Each channel consists of a detector microphone connected to a loudspeaker via an electronic controller. Using the method developed here to realise the controller it can be seen from figure 3.1.5 that the number of filter pairs needed in the controller is equal to the square of the number of channels; the complexity of the controller increases as the square of the number of channels thereby limiting the number of channels that can be practically implemented.

Consider a general control system where the number of monitors is greater than the number of speakers, which is a likely situation. In this case the term inside the brackets does not factorise because C is not square. However, by arranging the electronic feedback to cancel the acoustic feedback the term F can be set to zero to obtain the derired response of the feedforward filter;

$$K = -(C^H C)^{-1} C^H E$$

This expression for the feedforward path is valid for control systems consisting of any number of detectors, sources and monitors providing that the electronic and acoustic feedback paths match.

By using the topology indicated in figure 3.1.5 to realise the controllers the filter pair connecting each detector to each source is repeated a number of times. This is of great advantage when designing a controller; the reproducibility prevents each new controller needing to be redesigned from scratch. The filter pair consisting of a forward filter in parallel with a feedback filter can be used as the basic building block with which to realise all controllers. If the controller was to be implemented on a single microprocessor then the software forming the filter pair could be duplicated the desired number of times. Perhaps a more likely situation is that the basic unit of the controller be implemented on a single chip and to implement higher order controllers a number of chips could be used.

The first step in demonstrating the idea in practice is to implement and test the basic controller unit. The implementation of the basic controller unit on a Texas Instruments TMS32020 microprocessor is described next.

3.1.6 SUMMARY

The practical need for multichannel active noise controllers has been discussed and the equation giving the required responses of the controllers reviewed.

The controller responses were derived for specific cases and it was shown how the topology of the filters making up the controller needs to be considered. A new method was presented whereby, by using a specific arrangement of filters where electronic feedback paths are specifically designed to cancel the acoustic paths, the expression for the desired responses of the controller is simplified and the controller for any multichannel controller can be readily realised.

It was shown how any multichannel controller can be realised by repeatedly using the same filter pair as used in the single detector-single source system. This thesis proceeds by describing the practical implementation of this basic unit of a multichannel controller.

3.2 A PRACTICAL IMPLEMENTATION OF A DIGITAL CONTROLLER FOR A SINGLE CHANNEL ACTIVE NOISE CONTROL SYSTEM

3.2.1 INTRODUCTION

This section is concerned with the implementation of the digital controller for an active noise control system consisting of a single detector and a single source which could control the field at one or more positions. The digital controller is realised by a pair of finite impulse response filters, one in the forward direction to be positioned in the path from the detector to the source, and the other in a feedback path to cancel the acoustic feedback.

It was shown in section 3.1 how the controller for a multichannel system could be realised by repeatedly using the same filter pair as described above. This implementation of the single channel controller therefore tests the basic unit of a multichannel controller.

It is shown how the filter pair can be implemented using one or two digital systems; the methods are compared, the advantage of implementing the controller on a single microprocessor is highlighted and it is explained how such a system can be stable.

The hardware used and the additional hardware developed are presented (3.2.3). The implementation of the controller on a Texas Instruments TMS32020 microprocessor is described. The operation of the program is tested and particular attention is paid to the position of the filter coefficients and the convolutions within the program (3.2.4-3.2.7).

Finally, the development of a digital system to implement a two channel active noise controller is presented.

3.2.2 TWO METHODS OF IMPLEMENTING THE ELECTRONIC FEEDBACK COMPENSATOR

The next step in the verification of the ideas presented in section 3.1 was to implement a single detector, single source control system to show that the digital controller constructed in this manner could successfully operate a sound control system. If the single channel controller could be shown to be successful then, in theory, a multichannel controller could be implemented by repeatedly using the digital system for the single channel controller.

It was required to implement a controller of the form shown in figure 3.2.1 where the feedforward filter connects the detector microphone to the control speaker and the feedback filter cancels the acoustic feedback.

This sub-section presents two methods of implementing the configuration of filters needed. It is explained how each method can be inherently stable and a suitable method is chosen to be implemented in practice.

The digital controller is essentially an infinite impulse response filter because the feedback results in a closed, therefore infinite loop. For this reason the electronic filter itself may be unstable; it will be absolutely stable however when the electronic feedback path is cancelling the acoustic feedback path, therefore the electronic feedback path would seem to need to match the acoustic feedback at all frequencies to ensure stability. However this is not practically possible because the the digital filter has a finite bandwidth. It is however only necessary to ensure that the digital filter gives a good match over its whole frequency range and that outside the working range the level is sufficiently attenuated in the electronic path. The question then becomes one of how to appropriately band limit the digital feedback filter to ensure good system behaviour outside the working range. This subsection presents two methods of implementing the digital feedback filter.

The digital feedback filter can be band limited with analogue low pass filters at entrance to and exit from the system. The first to remove high frequency components in a signal entering the digital system and the second filter to remove the high frequency components present in the sharp discontinuities in the digital signal at exit. The feedforward and feedback paths can be split so that each digital filter uses a pair of analogue filters. The complete control system then takes the form shown in figure 3.2.2. Let the following symbols refer to the following transfer functions;

F : acoustic feedback path to be cancelled.

H : combined transfer function of the two low pass filters.

To cancel the acoustic feedback F, the digital feedback filter needs to have a frequency response given by F/H. The disadvantage of this method is that the electronic filter required is not causal (see section 2.2); it needs to model the inverse transfer function of the analogue filters.

An alternative method of implementing the control system is shown in figure 3.2.3. This is the method that was practically implemented and presented in this thesis. Both the feedforward and feedback paths are combined in the same sampled data system; ie. implemented on the same microprocessor. In this case the desired frequency response of the feedback filter is simply F, the acoustic feedback itself from the exit to the entry of the digital system.

The obvious advantage in implementing the controller this way is that the system needs only one set of analogue filters, analogue to digital and digital to analogue converters and one microprocessor chip. In addition, the digital feedback filter only needs to model a causal path as opposed to the first method. These advantages heavily outweigh the fact that the feedback filter in the second system may need to be longer than in the first case because it incorporates the response of the analogue filters.

The digital system of figure 3.2.3 forms a closed loop and as previously presented, to ensure stability it is necessary that the system is band limited. A digital system sampling at F Hz has frequency components defined up to F/2 Hz (they will only be correctly defined if an infinite time length record is analysed). The spectrum is the result of a discrete Fourier transform on sampled data and therefore the spectrum is repeated at intervals of F/2 Hz. The response of the digital system is therefore defined over all frequencies. However, attention need only be paid to the response of the digital system up to a certain frequency (half the sampling frequency); if the closed loop of the digital system is stable over this frequency range then it is stable over all frequencies because the response at higher frequencies is simply a repetition of the response at a lower freqency.

This explains how the closed loop digital filter configuration can be stable. It is also necessary that the complete control system be stable. This can only be achieved if the match between the digital feddback and the acoustic feedback is accounted for over the whole frequency spectrum. If the digital filter is long enough to sufficiently model the acoustic feedback then the match can be achieved over the working range of the digital system (ie. up to the nyquist frequency) (see section 2.2).

This explains how the closed loop digital filter configuration can be stable. For stability the fall off of the analogue filter at exit from the digital system needs to be sharp enough to attenuate the high frequency components sufficiently.

3.2.3 THE HARDWARE USED IN THE PRACTICAL IMPLEMENTATION AND THE ADDITIONAL HARDWARE DEVELOPED

The hardware presented in this sub-section was used both to implement the digital controller and record the necessary acoustic responses in the experiments described in chapter four.

The digital control system was realised using a Texas Instruments TMS32020 microprocessor. The chip has a single cycle multiply and accumulate instruction using 32 bit arithmetic making it ideal for convolving two time series thereby implementing FIR filters. The microprocessor was mounted on a board from Loughborough Sound Images Ltd, housed in a Ferranti PC860XT personal computer. The same system was also used to record measurements on the acoustic system needed to derive the digital filters (sections 3.3 & 4.2). Additional hardware was constructed to interface to the digital system.

The TMS320 board incorporated an analogue to digital converter and a digital to analogue converter. An ADC functions by converting the analogue signal at the input to a quantised digital value. However, the conversion is not instantaneous and the voltage level presented to the input of the ADC needs to be held constant during the conversion. The conversion time of the ADC was given as 17 microseconds (Ref. TMS 32020 board User Manual). Over this time the signal presented to the input can change. At 300 Hz the fraction of a cycle that the signal can change by is given by; 0.000017 x 300 = 0.0051 ie. about 2 degrees of a cycle which can introduce a significant error at the lower signal levels. Therefore the analogue value presented at the input to the ADC needed to be held constant while the ADC was quantising the signal. This can be achieved by holding the voltage level constant with a sample and hold device clocked at a suitable rate.

This thesis also describes the implementation of a digital controller for operation with a control system consisting of a single detector and two control sources. This required the output from the DAC to be demultiplexed between two channels; ie. for the single DAC to serve two output channels. It was also considered that in future work in this laboratory the processor may be required to operate with two inputs; for example when running an adaptive filter system which would require another input to generate an error signal aswell as requiring the signal which was to be filtered. For this reason an interface to the TMS32020 board was designed and constructed to input and output two channels. This electronic device needed to be purpose designed to operate with the TMS32020 board. The technical specifications of this multiplexer - demultiplexer circuit are covered in appendix 1.

The multiplexer/demultiplexer unit (MDMU) was designed and constructed to be operated by a specific set of instructions. The control logic could be operated in a variety of ways to sample or output however many signals are required in a clock cycle and at the required instant depending on the task at hand. However, because the microprocessor operates on sampled data then it is always necessary to sample and output the data points at a chosen clock rate. The control logic is somewhat complicated and altering the logic for a different task would be a tricky operation. Therefore the MDMU was controlled with a specific set of instructions which caused two samples to be input to the system and two samples to be output (one on each channel) on each cycle. The output signals have previously passed to the unit via the DAC and the input signals are subsequently passed to the ADC. The unit and the software routine which controls it can be regarded as a stand alone system. Any program to be implemented on the system can conveniently be inserted into this software controlling the inputs and outputs to the digital system.

This section proceeds by describing those elements of the digital controller particular to the filtering operations required in the controller program.

3.2.4 IMPLEMENTATION OF THE DIGITAL CONTROLLER ON A TEXAS INSTRUMENTS TMS32020 MICROPROCESSOR

The next four sub-sections describe the operation and testing of the software used to implement the digital control system. This sub-section explains the organisation and flow of the control filter program.

The implementation of the digital controller involved writing a program in assembly language to run on the TMS32020 microprocessor. The program realised two FIR filters configured as in figure 3.3.1 and is listed in appendix 2. The processor has an on chip memory (the memory which is easily accessible) of 544 words, 512 of which are stored in two blocks such that the blocks can be convolved easily (Ref. Texas TMS32020 User Guide). To implement the two filters it was necessary to store two sets of filter coefficients and the input and output time series of the same length. Therefore the 512 words of on-chip memory (known as data memory) were equally divided between the four time series that needed to be stored. This resulted in filter lengths of 128 points. A diagram of how the main data storage blocks were used is shown in figure 3.2.4.

The program listed in appendix 2 shows that the coefficients are listed as part of the program itself and are therefore stored as part of the program memory when the program is loaded to the microprocessor. A new set of coefficients could be loaded into the program as follows. The coefficients were stored separately as integers in a data file and a Fortran program read the assembly language program and the set of filter coefficients, combined the two and wrote a newly formed assembly language program. The coefficients listed in the program are the coefficients used in the active sound controller implemented in section 4.

What follows refers to the manner of operation of the program itself (appendix 2). The coefficients are first loaded from the program memory to the data memory block BO. The block is then configured as program memory to enable use of the instruction which performs the convolution. The MACD instruction (standing for multiply and data move) multiplies the contents of a data memory address with the contents of a program memory address in block BO. The result is added to the contents of a specific register and pointers automatically move to refer to the next locations to be multiplied making the instruction ideal for performing convolutions.

The sampling rate of the controller was set to 1 KHz, although this could easily be altered. The program was set to operate in an interrupt mode; ie. for much of the sample period the processor idles and on each clock cycle it jumps to perform the main program routine; the interrupt service routine. What follows is a list of the sequential set of operations performed in the interrupt service routine.

read in XN the current sample from channel A of the MDMU

XN = XN / 2

to avoid overflow if XN and FN are both large

XN = XN + FN

add FN, the feedback term

$$YN = \sum_{i=1}^{128} X_i \; C1_{(129-i)}$$

convolution of feedforward filter

YN = YN / 8

because forward filter coefficients are scaled by a factor of 0.125

CHANA = YN x 2

get result ready to be output. Note that the filters cannot be convolved together in the implementation because the result of the convolution sum from the first filter is needed. The output time series from the feedforward filter is used as the input to the feedback filter.

$$FN = \sum_{i=1}^{128} Y_i \; C2_{(129-i)}$$

convolution of feedback filter

output CHANA

idle until the next clock pulse

At this time the program returns to the top of the above list to read in the next input value. The cycle operates continuously until the processor is reset.

It was shown in section 2.3 how, for a control system operating in a reverberant field the magnitude of the gains of the feedforward and feedback filters will be inversely proportional to each other; if one gain is of the order of n the other gain will be of the order of 1/n. In practice (section 4.2) the system was designed such that the feedforward and feedback paths both had gains of the order of one. It was found in practice that a gain of the order of unity in the feedback path resulted in a maximum coefficient of about 0.1 in that path. However, the maximum coefficient values in the feedforward path were typically greater than one and to implement this filter it was necessary to divide the coefficients by a scale factor (chosen as 8) and therefore necessary to scale the product of the forward convolution by the same scale factor.

The filter coefficients are positioned in program memory in reverse order. For example in the case of the feedback filter the larger coefficients at the start of the impulse response are positioned in the higher program memory locations. This can be seen in the increase in magnitude of the coefficients towards the end of the list.

The coefficients are loaded into the data memory in the same order with the start of the impulse response at the higher data memory locations. The feedback coefficients are convolved with the time series in the manner shown in figure 3.2.5. Let the current value of Y be Y_1; ie. the result of the feedforward convolution sum in this sample period. Let the value of Y from 127 samples ago be Y_{128}. The product of the convolution is given by the sum shown in the list above. $C2_1$ is the first coefficient in the impulse response of the filter. The feedforward filter operates in a similar way. It can be seen how the convolution occurs by visualising the progression of the time series over time; the values of Y are moved along the impulse response of the filter. A particuar value of Y is multiplied with the first value of the impulse response first and progresses along to the end of the filter being multiplied by the successive coefficients of the filter in each sample period.

3.2.5 CHECKING THE OPERATION OF THE DIGITAL CONTROLLER

The following checks were made on the operation of the program to determine that the coefficients were in the correct places and operating in the desired manner.

It was observed that the coefficients were correctly transfered from the program memory to the correct locations in the data memory.

The closed loop of the program was broken and checks were made with the output signal as YN, the result of the first filter convolution and the output signal as FN, the feedback term resulting from the input signal passing through both filters. Tests were made with short length filters which had definable responses.

A single coefficient equivalent to unity was loaded into the first position of a filter and the remainder of the values set to zero. A check was made that the transfer function between the input and output signals had an amplitude of unity and a phase delay of one clock cycle. The transfer function was viewed on a Hewlett Packard spectrum analyser and needed, as always, to be measured with low pass analogue filters at entrance to and the exit from the digital system. The phase delay through the digital system could only be measured by calculating the difference in the phase delay of the digital system and analogue filters and the delay through the analogue filters alone. It is interesting to note that in measuring the timing through the system in this way causes the digital system itself to result in an effective time delay of one and a half sample periods (ie. 1.5 mS at 1 KHz). This phenomenon is illustrated in figure 3.2.6 where the effective input and output waveforms resulting from a filter of a single coefficient are shown. The signal levels of the input and output points are shown. However the signal levels at the output are held at that value for the whole cycle; the signal passes through the reconstruction filter to the spectrum analyser which effectively records a voltage level as occuring half way along a sample period.

The position of the first coefficient in the filters was determined by the position at which a single non-zero coefficient gave a delay of only one cycle (ignoring the extra half cycle explained above). Positioning the pulse at the n'th coefficient position gave a delay of n+1 cycles through the digital system.

3.2.6 TESTING THE FIR FILTERS BY MODELLING A SIMPLE SYSTEM

An experiment was performed to verify the success of the filter program in modelling a simple system. The method of modelling and the necessary measurements used here are covered more thoroughly in section 3.3. The test was carried out to verify that the measurement procedure, analysis and filter implementation were compatible in their manipulation of the time series and filter coefficients.

An analogue device with a time delay of 20 mS (measured by recording the phase delay across it on a spectrum analyser) was chosen to be modelled. The test involved designing a digital filter so that the combined effect of the digital filter and the analogue filters would cancel the time delay (as shown in figure 3.2.7). The time delay was much longer than the group delay of the analogue filters thereby dominating the system to be modelled. This experiment is also of interest to the study of active sound control systems as, to a first approximation a time delay can model the delay down a duct. Previous experiments in this laboratory (Ref. Gurrie) had shown that a good match, in terms of the impulse response of the electronic path and the attenuation achievable can only be implemented when the time delay is greater than the delay through the analogue system. Therefore with such a system it can easily be determined if the time delay was being modelled adequately and the digital controller was operating in the intended fashion.

Let the transfer function through the analogue filters be H and the transfer function of the digital filter and the time delay be H1 and T respectively. Refering to figure 3.2.7 it is required that;

$$H1 = T / H$$

The digital system was used to pass a broad band swept sine signal through a time delay (where 'swept' refers to the frequency sweep of the signal). The signal x_2 was passed from the digital system through the time delay via the low pass filters and recaptured on the digital system. The spectrum of the resultant signal is given by;

$$H\ T\ X_2$$

The same swept sine signal was passed from the digital system through the analogue filters and recaptured. This process was repeated with the captured signal resulting in a signal which had passed through both sets of analogue filters twice. The spectrum of this signal is given by;

$$H\ H\ X_2$$

The ratio of these two spectra is given by T/H (figure 3.2.8). This represents the desired response of the digital filter. The amplitude spectrum is flat over the frequency range up to the cut off frequency of the low pass filters and the phase spectrum shows a linear delay of 4 cycles in 287 Hz; ie. a delay of the order of 14 mS. The inverse Fourier transform of this spectrum is shown in figure 3.2.9. Again the delay through the digital filter can be estimated by noting that the largest coefficients of the filter are of the order of 14 mS along the time axis. The missing 6 mS can be accounted for from the group delay through the analogue filters and a sample delay through the digital system itself. The impulse response of figure 3.2.9 is not a pure time delay because the digital filter has to compensate for the frequency response of the analogue filters; it needs to model the inverse response of the filters and the non-causal part of the impulse response resulting from trying to model a non-causal system can be seen at the right hand side of the impulse response; this effect is refered to further in the results of section 4.3.

A 128 coefficient finite impulse response filter was derived by deconvolving the two signals (figure 3.2.10). The filter was loaded into the feedforward filter of the program and the circuit of figure 3.2.7 implemented by breaking the closed loop of the program. The transfer functions of both the digital system and the time delay itself were measured and compared: it was verified that the electronic path matched the time delay. Similar measurements were verified when the filter was implemented in the position of the feedback filter of the program, with a single coefficient of unity magnitude in the first position of the forward filter and the result of the second convolution being used as the output from the digital system.

This test showed that the coefficients were correctly negated resulting in cancelling the path which was being modelled. The test also verified that the convolutions were operating in the intended manner in the program.

The convolution results in a value YN which has been calculated from XN and previous input values. However, YN is output one clock cycle later than XN is input because the system is operating in sample delay mode. It was important to be aware of this when implementing the system. However, to compensate for this a similar delay was present in the recording of the measurements used to derive the filters. What was practically important was that the measurement procedure and the filter implementation were compatible in their manipulation of the time series and the procedure resulted in the successful modelling of a system.

The previous test which modelled the time delay device confirmed that the filter coefficients were correctly positioned in each filter but did not test how the filters operated together in the closed loop.

3.2.7 TESTING THE CLOSED LOOP OF THE CONTROLLER USING FILTERS OF SIMPLE KNOWN RESPONSES

The operation of the control program has been explained and its practical testing described. However, the fully working controller consisted of a closed loop round the two digital filters. It was necessary to devise a test to verify the operation of this closed loop in practice. This needed to be achieved without breaking the loop and thereby altering the program. This sub-section describes the operation of the closed loop of the digital system and then proceeds to test the operation of the loop in practice.

The displacement of the coefficients in the digital feedback system also needed to be considered. If the coefficients were implemented as derived then the feedback term added to the input term would be one sample out of synchronisation. The current feedback term is the result of convolving the time series containing the previous, and not the current input sample.

The reason this phenomenon occurs in the feedback filter but not in the feedforward filter may be understood from the following; there is a sample delay through the feedforward path because the path passes through the input and output of the digital system between which there is a delay of one sample period. The digital feedback path is contained wholly within the digital system and does not incorporate the sample delay because the processing itself occurs in a very short time. The sample delay which is inherently compensated for in the measurement system is only needed in the forward filter and is not needed in the feedback filter. Therefore it is required to time advance the signal passing through the feedback path by one sample; this was achieved by shifting the filter coefficients by one place forward in time; ie. to the right in figure 3.2.5. Effectively the first coefficient of the feedback filter derived from deconvolving the two measured time series was ignored. (Alternatively the situation could have been resolved by deconvolving the two time series with the input time series advanced by one point). This theory was subsequently verified in practice when the feedback match could be upset by displacing the coefficients one place in either direction.

Tests presented so far have involved operating on the input and output signals from the controller program with the closed loop broken by not adding the feedback term to the input value. It was necessary to consider the two types of instability that could occur in this closed loop digital system. Firstly, a numerical overflow could occur in the convolution, and maybe not the result of the convolution itself but during the accumulation of the total as the sum alternates between positive and negative values. The scaling of the coefficients and samples prevented this. Secondly, the closed loop of the program needed to be stable.

Although in practice it was difficult to verify the successful operation of the closed loop program what follows is a test that indicated the closed loop functioned as required. A single coefficient of unity magnitude was positioned at the start of the feedforward filter and a single coefficient of magnitude one half was positioned at the start of the feedback filter. The resultant system is shown in figure 3.2.11. For an input value of X the current value of x is given by;

$$x = X + f\ e^{jwT}$$

$$x = X + 0.5\ y\ e^{jwT}$$

where T is the sample period ie. the current value of x results from the input value and the feedback term resulting from convolutions occuring in the cycle before. The output from the system, Y is equal to the value of y calculated in this cycle which is simply the current value of x multiplied by the filter coefficient of unity;

$$Y = x$$

$$Y = X + 0.5\ Y\ e^{jwT}$$

rearranging the equation results in;

$$X = Y\ (1 - 0.5\ e^{jwT})$$

The transfer function has a complex gain depending on frequency. At higher frequencies the gain falls off. It can be calculated at what frequency the magnitude gain equals one. This occurs when;

$$1 - 0.5\ e^{jwT} = 1$$

$$1 - 0.5 \cos wT - 0.5\ j \sin wT = 1$$

$$\cos w/1000 = 0.25$$

ie. the magnitudes of the input and output are equal at 210 Hz. This was confirmed in practice (figure 3.2.12).

This test using simple definable impulse responses in each filter confirmed that both filters were correctly positioned relative to each other. The transfer function across the controller agreed with theory indicating that the closed loop of the program was acting successfully.

3.2.8 IMPLEMENTATION OF A DIGITAL CONTROLLER FOR A TWO CHANNEL ACTIVE NOISE CONTROL SYSTEM

Having developed an assembly language program to control a single channel active noise control system the technical methods used to implement the system were duplicated to implement a controller for a two channel system. The digital controller developed could be used to operate an active sound control system consisting of a single detector microphone and two control loudspeakers capable of controlling the field at two locations. Such a system would not be used to control the field at only one location because this would overspecify the controller required. The assembly language program for a two channel system is listed in appendix 4.

The size of the memory of the TMS32020 microprocessor and the sampling rate required limited the number of FIR filters that could be implemented on a single chip. The two channel controller needed four filters configured as in figure 3.1.4. As well as storing four sets of coefficients, the input time series and two output time series needed to be stored; seven data sets in all. The controller was to operate in the same environment and therefore it was desired to implement filters of the same length as the single channel controller (ie. 128 points). The on-chip memory of the microprocessor was only 512 words so some data transfer operations needed to occur during the program cycle. As in the single channel controller the input time series was stored in page seven of the memory (figure 3.2.4). The filtering operations of each channel were performed in turn. Before the convolution can take place both sets of filter coefficients are loaded from program memory to the data memory block B0 where they need to be for the convolution. The output time series for that channel, also stored in program memory, is loaded into the data memory block B1 (page 6). After the convolution, the output time series, which has shifted by one point is returned to be stored in the program memory. The coefficients are not shifted and do not need to be restored. The same operations need to be performed on the other channel of the controller. The time required for thse data movement operations means that the maximum clock rate of the program is 3.7 KHz; operating at a quicker rate the processor does not have sufficient time to complete all the required instructions.

The two channel controller was implemented after a single channel control system had been successfully implemented (chapter 4). The filter coefficients derived in section 4.2 could be used successfully with the input and output to the system through either channel. This demonstrated the successful operation of both channels of the multiplexer/demultiplexer unit. The single channel controller could also be operated with the two channel controller program operating through either channel. This verified the successful operation of the two channel controller program.

3.2.9 SUMMARY

It has been discussed how it is necessary to band limit the electronic feedback path of the controller to obtain a suitable match with the acoustic feedback path. It has been shown how it is possible to realise both the feedforward and feedback filters of the controller on a single digital system and that the system can be stable.

The implementation of the controller on a TMS32020 microprocessor has been described. It was shown how the sampling of the signals and the convolutions need to be carefully considered in the implementation.

The controller program was successfully used to model the response of an analogue time delay device and the frequency response of the controller containing simple definable filters was shown to agree with theory. These tests demonstrated that the controller was operating as intended. Finally the practical implementation of a digital controller implemented for use with a two channel active control system was described.

Having developed a practical method of implementing a control system the next section describes a method used to derive the filter coefficients needed for a particular controller.

3.3 A PRACTICAL METHOD OF DERIVING THE DESIRED CONTROLLER

3.3.1 INTRODUCTION

The previous section described the practical implementation of a digital controller for a single channel active sound control system. Having developed the particular digital controller the digital filters for a given situation need to be evaluated.

A practical method is presented to derive the digital control filters for an active controller consisting of a single detector and a single source controlling the field at one position. The method is used to experimentally derive the digital filters in the practical implementation of chapter 4 using the digital controller described in section 3.2.

The response required for the digital controller is reviewed and a need is shown to measure various transfer functions in the system. The aims and requirements of a method to do this are discussed and an appropriate method presented.

The measurements used to record the system responses are defined and the experimental details of the practical method used to record the responses are presented. The measurements need to be analysed to derive the digital filters; also the method used in practice to do this is presented.

Finally, a practical method is developed to quantify the stability (or instability) of the control system. It is shown how the controller of section 3.2 and the practical method used to characterise the system were combined to form a system to record the open loop transfer function of the controller.

<u>3.3.2 REVIEW OF THE FILTER STRUCTURE USED TO REALISE THE CONTROLLER</u>

The previous section described how the digital controller was practically realised; by using two finite impulse response filters which have fixed impulse responses. The requirement of the filters is that they have a frequency response particular to the response of the control system and the environment it operates in.

The frequency response of a filter can be realised digitally in a number of ways, using filters with finite or infinite impulse responses (Ref. signal processing texts). Ideally it is desired that a control system would work adaptively, continually updating to changes in the noise source and environment. Adaptive filters and IIR filters both operate by having feedback in the filter structure. This feedback can cause instability; although works have been published describing the use of adaptive digital filters in active noise control systems (Ref. Erikson, Poole) employing these filtering methods is not straightforward and the most appropriate methods have yet to be defined. The most straightfoward method of digital filtering employs finite impulse response filters. This thesis describes the successful implementation of a controller using FIR filters which have fixed impulse responses.

The filter responses are determined by the responses of other paths through the system (see section 2.2), therefore it is required to measure these responses by some method and process the measurements to derive the control filters.

3.3.3 AIMS AND REQUIREMENTS OF MEASUREMENTS NEEDED TO CHARACTERISE THE CONTROLLER

The response of the controller needs to be defined over the frequency range to be controlled. Therefore it is necessary to measure the system responses over a frequency range which encompasses the desired working range of the control system. Beyond this, however, it is necessary to ensure that the system does not cause enhanced noise levels at outlying frequencies and that the system is stable. Instabilities often occur at high frequencies outside the working range, demonstrated by the howl sometimes heard in acoustic systems, so it is necessary to specify the system response at all frequencies.

The digital control system is band limited by the low pass filters at the entrance to and exit from the system; therefore it is necessary to measure the system responses up to and beyond the cut off frequency of the low pass filters, to a frequency where components are no longer passed significantly.

It is important to use appropriate equipment to record the system responses. System responses can be conventionally measured on a spectrum analyser. However, the intention of measuring these responses is to generate a digital filter which compensates for the environment it is working in; this includes the electronic environment as well. Therefore it is desirable to use the same hardware for measurement as is to be used for the control system, thereby compensating for the response of the electronic measurement system itself. Consider the action of a two channel Fourier transform spectrum analyser. It takes a length of sampled data and transforms the time series into a frequency spectrum (perhaps using a fast Fourier transform). A system transfer function between input and output signals is calculated by dividing the two complex spectra. Now if the impulse response of the system is required this spectrum needs to undergo an inverse transformation. A transformation is only accurate for all frequencies if an infinite time length is used and transforming both ways could lose information contained in the time series and distort the result. The transformation will be most inaccurate at low frequencies where the time period approaches that of the record length.

As the intended result of this system analysis is a time series (the impulse response of the digital filter) it is desirable to process the data wholly in the time domain.

The first lobe of the spectrum of a rectangular pulse of length T has a frequency width of 1/(2 T) and if we are concerned with defining frequency responses of the order of 250 Hz a pulse of less than 2 mS long would be needed. It is difficult to measure responses with very short duration pulses, there is often not enough energy in such a pulse to excite a system and signal to noise ratios are very small because only a small amount of energy is present in each frequency component.

The filter whose response is to be measured needs to be excited by a suitable signal. This signal needs to be defined in frequency up to and beyond the upper frequency of interest; in this case beyond the cut off frequency of the low pass filters. It would be possible to use a random noise source with a suitable frequency range but the measurement can be performed much more quickly and efficiently using a transient signal sweeping the desired frequency range. This method has previously been used successfully in acoustical studies in this laboratory (Ref. Howes, Chapman). This method has the advantage that the spectrum of the signal is well defined for whatever frequency range is required and can be arranged to have a relatively constant amplitude thereby representing all frequencies equally.

The acoustical measurements are carried out using a transient signal; this method is appropriate because the active control system is not concerned with quasi steady state conditions. The response time of a system is determined by the amount of damping in the system. The question could be posed as to what is the maximum rate of frequency sweep of the transient signal, such that a resonance is excited for long enough to attain maximum amplitude. However this is only of interest when measuring the response under quasi steady state conditions. In the case of the active control system the resultant controller will track the response of an acoustic system and prevent the resonances being excited to their full extent; quasi steady state conditions are not reached, nor intended to be.

In summary, it has been discussed how the response of the controller needs to be defined over all frequencies. Therefore the measurements used to characterise the system and derive the filters need to be defined over the working range of the system. It has been shown why the measurements used to characterise the system need to be performed with the same hardware used to implement the controller. It is preferable to perform the analysis in the time domain and a suitable method of measuring the system frequency responses is presented.

3.3.4 A SERIES OF MEASUREMENTS TO CHARACTERISE THE CONTROL SYSTEM

The method used to measure the system responses was as follows; The digital system used as the controller was used to excite the acoustic system whose response was to be measured. The response was also captured on the digital system. Before presenting the practical details of this measurement method, the measurements used to characterise the acoustic system and derive the digital filters are presented.

Previous work at RHBNC (Ref. Gurrie) has established a series of acoustic measurements from which the digital controllers can be derived. The procedure involves exciting the system with a transient swept sine signal and synchronously capturing the response.

A signal was numerically generated using a Fortran program (Appendix 4) in which the frequency limits, the sampling rate and the number of points in the signal could be chosen. The signal was of the form

$$x(t) = x0 \sin((a + b\ t)\ t)$$

where b is the angular frequency range of the sweep divided by twice the time length of the sweep;

$$b = (wu-wl)/(2\ T)$$

where wu and wl are the upper and lower angular frequencies of the sweep and a is the angular frequency of the start of the sweep;

$$a = w1$$

The instantaneous frequency of the signal at a time t is given by the rate of change of the angle inside the sin term.

$$w = \frac{d}{dt} ((a + b\ t)\ t)$$

$$w = a + 2\ b\ t$$

Hence the frequency of the signal linearly increases from the lower frequency a to a frequency a + 2 b T

In what follows capital letters denote SPECTRA and small letters are used for signals as functions in time.

Consider the control system acting in the enclosure (figure 3.3.1). The relevant transfer functions are shown in the diagram and the notation is the same as that used for the duct system in section 2.2.

The desired response of the feedback filter is F and the desired response of the feedforward filter is −E/C (where E = A/B) = −A/(B C)

The response of the acoustic feedback path is relatively easy to measure; let the swept sine signal x_2 injected into the control speaker result in a signal y_{12} (y denoting an output signal, $_1$ and $_2$ denoting points in figure 3.3.1). The spectrum Y_{12} is the product of the spectrum X_2 and the transfer function F. Therefore the response of the acoustic feedback path is given by;

$$F = Y_{10} / X_2$$

This response is the desired response of the digital feedback filter.

The method of measuring the feedforward filter is slightly more complicated. The swept sine signal x_2 injected at point O to excite the noise source loudspeaker results in responses y_{10} at the detector microphone and y_{30} at the monitor microphone. The transfer function between the two microphones is given by;

$$E = Y_{30} / Y_{10}$$

The transfer function C is the ratio of the spectrum of a broadband signal exciting the control source and the response at the monitor microphone. However, consider the action of the digital controller; it only receives those frequency components which have been detected at the monitor; only these frequency components proceed to excite the control speaker. Therefore it is valid to excite the control speaker with the signal previously captured at the detector (y_{10}). Denote the response at the monitor as y_{32} and the transfer function C is given by;

$$C = Y_{32} / Y_{10}$$

The response of the feedforward filter is given by;

$$-\frac{E}{C} = -\frac{Y_{30}}{Y_{10}}\,\frac{Y_{10}}{Y_{32}} = -\frac{Y_{30}}{Y_{32}}$$

Conveniently the spectrum Y_{10} has dropped out of the analysis leaving the filter response given by the ratio of two spectra.

In summary, the following measurements are recorded;

A transient swept sine signal (x_2) excites the noise source loudspeaker and the responses captured at the detector microphone (y_{10}) and monitor microphone (y_{30}).

The signal y_{10} is used to excite the control speaker and the response captured at the monitor microphone (y_{32}).

The transient swept sine signal is used to excite the control speaker and the response captured at the detector microphone (y_{12}).

This section proceeds by describing the practical procedure developed to record these system responses.

3.3.5 A PRACTICAL METHOD OF RECORDING THE RESPONSES OF THE CONTROL SYSTEM

This subsection describes the practical method developed to record the time responses of the acoustic system. The method presented here is applicable for measuring the response of any device or system. The digital system used for the measurements was the same system on which the controller was implemented (section 3.2).

The swept sine signal used as the excitation signal was numerically generated by a Fortran program (appendix 4). The mathematical form of the signal determined that the instantaneous frequency of the signal increased linearly between two frequency limits. The range of the frequency sweep could be specified by the user of the program. The sampling rate, the number of points in the signal and the length of signal to be windowed at each end (to reduce sharp discontinuties) are also specified by the operator. The more precise requirements of the practical signal used are covered in greater detail in section 4.2.

At this stage the development of a suitable excitation signal has been described. What follows is a description of the practical method used to record the response of the system under test.

The swept sine signal was downloaded to the TMS32020 by converting the numerical data points into Texas object code (see Ref. Texas) and loading the object code into the program memory of the microprocessor.

The structure of the program used to excite a system and record the response is quite simple; on each clock cycle a data point of the output signal (the swept sine signal) is output from the digital system and the input channel to the digital system is sampled.

It was explained in section 3.2 how the digital system (the TMS32020 and interface electronics) was adapted to input and output two channels on each clock cycle, irrespective of the program in operation. The input and output operations peculiar to the digital system are covered in appendix 1.

The assembly language program used to excite a system with a stored signal and capture the system response referred to in the text as the data capture program is listed in Appendix 5. The program outputs and inputs 1900 data points. The output signal is initially stored in program memory (off-chip) and output via a specific data memory location (on-chip). The input signal is input via a data memory location to be stored in the program memory.

The operation of outputting and inputing 1900 data points is repeated sixteen times and the sampled responses averaged. This averaging was intended to increase the signal to noise ratio of the captured signal. The averaging is achieved by dividing each input sample by sixteen and summing the signals. Initially zero is stored in the program memory locations where the signal is to be stored. On each clock cycle a data point is input, divided by sixteen and added to the current sum of that particular point. Because the dynamic range of each input point is reduced by a factor of sixteen the quantisation noise of the measurement is reduced. However, the overall signal to noise ratio is governed by that of the electronic input unit (see section 3.2 and appendix 1) and is of the order of 40 dB.

When the signal has been stored and the averaging is complete the program is able to repeatedly output the captured signal enabling it to be displayed on an oscilloscope until the processor is reset. This enables the user to check the form of the captured signal. If the signal were truncated, clipped or the measurement had been disturbed by an unwanted noise then the measurement could be repeated.

The captured signal could be uploaded to a data file on the host personal computer by saving the relevant section of the program memory.

Having obtained the response of a system to a broad band excitation the system responses can then be analysed. The report proceeds by presenting the signal analysis used to derive the required digital filters.

3.3.6 THE SYSTEM MODELLING USED TO DERIVE THE DIGITAL CONTROL FILTERS

The digital filters to be used in the control system are derived from an analysis of the system itself. The filters are numerical models of a desired system response; hence the term system modelling. The analysis determining the desired responses of the digital filters has been presented in terms of the spectra of the signals used to characterise the control system but, as discussed previously, it is desirable to operate in terms of the time signals themselves. Indeed, the raw data available from the measurements constitute a number of time series.

In the frequency domain the desired response of the feedforward filter is given by;

$$H_1 = Y_{30}/Y_{32}$$

and the response of the feedback filter is given by;

$$H_2 = Y_{10}/X_2$$

where the terms Y_{30}, Y_{32}, Y_{10} and X_2 are the spectra of the measurement signals. The end result of this analysis is to determine two time series; ie. the impulse responses of the filters. However, the result is not readily available in the time domain. Transforming the equations into the time domain gives;

$$y_{32} * h_1 = y_{30}$$

$$x_2 * h_2 = y_{10}$$

To obtain the impulse responses h_1 and h_2 it is necessary to deconvolve the time series y_{32} with y_{30} and x_2 with y_{10}.

Consider a system having input x and output y related by the convolution;

$$y = x * h$$

where each term is a sampled time series. The individual terms of the series y can be specified as;

$$y(1) = x(1)\ h(1) + n_1$$

$$y(2) = x(2)\ h(1) + x(1)\ h(2) + n_2$$

$$y(3) = x(3)\ h(1) + x(2)\ h(2) + x(1)\ h(3) + n_3$$

etc.

The convolution is only a linear numerical approximation to a real system. Therefore terms representing the non linearity and the system noise have been added. The result of the operation of deconvolving the series x and y is to obtain a set of values for the series h such that when convolved with x the result gives the best fit to the series y. It is desired to obtain a numerical approximation to the series h. There are a number of different methods to model the series (see texts on system modelling). The method used in practice was a least squares fit in the time domain. This method may not be the best method but it is one of the most straightforward to understand.

Let the result of the convolution between x and h be y' and let the output from the real system be y. A set of values of h can be derived by minimising the squared error between the measured system output, y and the finite impulse response filter output, y'. A Fortran program of a least squares system identification algorithm (Ref. Marple) was used to deconvolve the time signals, resulting in a least squares error approximation to the system impulse response.

The remainder of this sub-section details the particular procedure used to derive the FIR coefficients. The data capture program output and input 1900 data points in total. However, the package used to display the signals and their spectra (Ref. ILS) could only analyse signal lengths of 512 points maximum. It was found that this signal length was adequate to derive the digital filters. Therefore the captured signal response was limited to 512 points in length and input and output signal lengths of 512 points were used as input to the deconvolution program.

The length of h needs to be chosen to be long enough to model the impulse response of the system adaquately. Given this then the estimate of h will improve as the length of h decreases relative to the length of x and y. This can be seen by considering the situation of h being longer than x and y; in this case the simultaneous equations cannot be uniquely solved. As the length of h decreases with respect to x and y the excess of equations are more adequately able to define the set of h. For this reason it was an advantage to perform the analysis with time signals of significantly greater length than the filters to be derived.

The filter lengths were to be 128 points (see section 3.2) and the signal lengths from the measurement were 512 points long. However it was known that there was no input to nor output from the system before it was excited by the swept sine signal. Therefore, to increase the signal lengths used in the analysis, the deconvolution was performed with signal lengths of 1024 points, the first 512 of which were of zero level.

The operation of the program was the most time consuming part of the analysis of the control system, taking about twenty minutes to perform the deconvolution on the Ferranti personal computer.

It has been shown how a series of measurements can be performed on a single channel active control system from which the required digital controllers can be derived. The practical method developed to record these measurements has been described and the theory of the signal processing used to derive the flters has also been presented. The Fortran program of the least square system identification algorithm was used to derive the digital feedforward filter from the measurement signals y_{32} and y_{30}. The feedback filter was derived from deconvolving the signals y_{12} and x_2.

3.3.7 THE PRACTICAL METHOD IMPLEMENTED TO MEASURE THE OPEN LOOP TRANSFER FUNCTION OF THE SINGLE CHANNEL ACTIVE NOISE CONTROL SYSTEM

Having developed a digital system to implement the active controller and a method to derive the specific digital filters needed it was desirable to test the stability of the system; to find over what frequency range, if any, the system was likely to be (or was) unstable. As remarked in section 2.2 instability round a loop of a system occurs when the loop has a gain greater than or equal to one at a phase lag of an odd number of cycles. To measure the extent that the loop gain of a system approaches this it is necessary to break the loop and record the transfer function round the loop.

Consider the active noise control system shown in figure 3.3.2. The path through the digital system is shown broken. It is required to determine the complex loop gain round the system starting and ending at this point. Note that the loop does not include the monitor microphone which is not needed for the controller to operate (but is only needed for the derivation of the digital filters).

The loop is broken in the feedforward path of the digital controller. It is necessary to break the loop here because breaking the circuit in either the digital or acoustic feedback path would have resulted in a closed loop remaining in the circuit (the loop round the opposite feedback path to the one that was broken). If this loop was unstable then the instability would become apparent as a saturated signal level from that loop but the test would not indicate over what frequency ranges the loop was unstable.

Therefore to test the stability in this manner it was necessary to break the circuit at a place common to both feedback loops ie. along the digital feedforward path. Therefore the open loop frequency response of an active control system could not be readily measured with a spectrum analyser but a program had to be implemented on the digital system itself to break the loop in the forward path and record the complex gain round the loop. The symmetrical position of the break indicated that some useful measurements could be recorded; such as the match between the digital and acoustic feedback paths.

The method used to perform the necessary acoustic measurements on the control system involved exciting the system with a swept sine signal and synchronously capturing the system response. A similar measurement procedure was used to record the open loop transfer function of the system. The swept sine signal was used to excite the system at a point just before the feedforward filter (figure 3.3.2) and the resultant signal captured at the same place.

The program to perform these measurements was developed by adapting the controller program (see section 3.2) and incorporating into it the elements of the data capture program. The program is listed in appendix 6.

It was necessary to verify the operation of the program used to record the open loop response of the control system. The operation of the program was compared to the operation of the data capture program. The equivalent circuits of figure 3.3.3 were implemented and the swept sine signal used to excite the analogue low pass filters. The response of the filters was recorded using the data capture program and the same response measured using the open loop response program, by setting the feedback filter to zero and positioning a single pulse at the start of the feedforward filter. The amplitude and phase spectra of both signals were the same indicating that the programs operated in the same manner.

The internal operation of the program was tested by positioning a pulse at the start of each digital filter (figure 3.3.4) and recording the response round the open loop of the digital system alone. The transfer function round the loop had a magnitude of one and a delay of one clock cycle, exactly as it should have been.

This program is used in section 4.3 to record the responses of the controller implemented.

3.3.8 SUMMARY

It has been shown how the frequency response of the active sound control system needs to be characterised to derive the necessary digital filters to control the system. It was shown how it is desirable to measure the system responses with the same digital system on which the controller is to operate and to process the raw data in the time domain.

The development of a practical method to record measurements on the controller has been presented and the series of measurements used to characterise the control system were derived. It was shown how the measurements were analysed to derive the digital filters.

Finally, the practical method used to record the open loop response of the controller has been presented.

4. THE PRACTICAL IMPLEMENTATION OF AN ACTIVE SOUND CONTROL SYSTEM OPERATING ON THE REVERBERANT FIELD INSIDE A SMALL ENCLOSURE.

This chapter presents the practical work carried out (using the methods presented in sections 3.2 and 3.3) to support the theory of sections 2.3 and 3.1. The chapter is divided into four sections the first three of which are concerned with the implementation of a single channel active sound control system inside a small enclosure.

Section 1 describes the test conditions and apparatus used. Section 2 presents the acoustical measurements on the enclosure and the digital control filters derived from the measurements. Section 3 presents some practical measurements used to assess the degree of success of the control system and the practical attenuation achieved by the system. The first three sections can be viewed as a whole, presenting the results of a practical implementation of a single channel control system.

Section 4 concerns some additional experiments on the single channel control system (i) operating at a higher sampling rate and (ii) working to achieve attenuation at a different monitor position from the experiment forming the bulk of this chapter. A single detector, double source control system is also implemented and the results discussed.

4.1 TEST CONDITIONS

4.1.1 INTRODUCTION

This section describes the small enclosure under test, the test conditions, practical apparatus, the physical limitations present and how these were optimised to design an experimental situation where an active noise controller could be expected to work.

Preliminary experiments on the sound field in the box are described showing the suitability of the box for a demonstration of an active control system attenuating the reverberant field.

The major constraints of the electronic hardware are shown to determine the maximum length of finite impulse response filters that could be readily implemented. The working range and sampling frequency of the system are chosen. Finally the operating limitations of the controller are discussed and the transducer positions for the control system operating inside the enclosure are chosen and the reasons for these choices discussed.

4.1.2 TEST ENCLOSURE AND SOUND FIELD

A rectangular hollow box, sealed on all sides but with a detachable lid was used for the test enclosure. The box had dimensions 0.49 x 0.595 x 0.705 m and its walls were constructed of 1.3 cm thick stiff perspex. It was of a convenient size to be placed on a workbench and its aspect ratio was such that the modal frequencies of the first few low order modes were distinct.

It was observed how the sound field inside the box was made up of contributions from the resonant modes of vibration of the air inside the enclosure. The modal frequencies agreed with the values predicted by the theory of section 2.1 to within 15%, suggesting that the theoretical equation predicting the pressure at a point from the mode shapes (section 2.1) was a good representation of the sound field inside the enclosure.

Representative plots of the sound field inside the enclosure are shown in figure 4.1.1. The field was excited by band limited pseudo random white noise driving a Bower & Wilkins mid-range loudspeaker (MR100) placed in a corner of the enclosure; a position that couples well into all modes. The transfer function between input to the noise source and the signal at the microphone was recorded on a Hewlett Packard HP3582A spectrum analyser.

The plots indicate the increase in sound pressure level towards a corner of the enclosure; this is due to the modal form of the sound field. It can be seen how there are no modes noticeable in the amplitude spectrum at the centre of the enclosure; the microphone here was positioned on a node of the low order modes within this frequency range (the 1,0,0 and 0,1,0 modes). A similar pattern is seen in the phase plots, large phase shifts occuring around the resonance frequencies but not present in the phase response recorded at the centre of the enclosure.

The apparatus of figure 4.1.2 was used to investigate if suitable levels of attenuation could be achieved inside the enclosure to make an active noise control demonstration worthwhile in this environment.

The sound field was excited using a loudspeaker positioned in one corner of the enclosure and a second speaker driven from the same noise source via a phase shifter was used to cancel the sound field (figure 4.1.2). The degree of cancellation achieved was recorded on a microphone placed at various positions throughout the enclosure. Two experiments were performed to test the suitability of the enclosure.

Firstly it was observed how the field dominated by a single mode could be cancelled with a single speaker. The first low order mode was excited at the modal frequency with two sources driven from the same oscillator (figure 4.1.2); the phase shifter could be adjusted so that the mode could be significantly attenuated with the cancelling speaker in a variety of positions.

Secondly a noise source of 100 Hz bandwidth, centred at 250 Hz was used to excite the first two low order modes. With the speakers placed as near as was practically possible (centres about 10 cm apart) the maximum attenuation measured was 8 dB at a point distant from the speakers in the enclosure. This was achieved by driving the two speakers from the same oscillator and reversing the terminals to one of the speakers. Similar attenuation levels could be measured throughout the enclosure. However, with the cancelling speaker in an opposite corner broadband attenuation could not be achieved. In this case the direct field could no longer hope to be cancelled, but cancellation could be achieved over a narrow frquency range centred round the mode resonances.

These preliminary demonstrations showed that the direct field could only be attenuated by placing a contol speaker near to the noise source. The reverberant field could be attenuated by a loudspeaker in other positions inside the enclosure.

4.1.3 WORKING RANGE AND SAMPLING FREQUENCY OF THE DIGITAL SYSTEM

A diagram of the control system showing the apparatus used is shown in figure 4.1.4. Some of the apparatus can be seen in the picture of figure 4.1.3 and the remainder is shown in the picture in figure 4.1.5.

The first two modes of the enclosure had resonance frequencies at 246 and 292 Hz. The two low pass filters at entrance to and exit from the digital system determine the working frequency range of the system. The cut off frequency of the low pass filters was chosen to be 350 Hz to encompass the two low order modes of the enclosure.

The sampling theorem states that only frequencies up to the Nyquist frequency can be detected in a sampled time series. Frequencies above the Nyquist frequency will be aliased as reflections about the Nyquist frequency. In order that frequency components beyond the Nyquist frequency would be sufficiently attenuated it was important that the sampling frequency was greater than twice the cut-off frequency of the low pass filters. For this reason the sampling frequency was chosen to be 1 KHz.

4.1.4 IMPULSE RESPONSE OF THE REVERBERANT SOUND FIELD

When modelling any system with a finite impulse response digital filter it is important to know the time length of the impulse response of the system to be modelled. In this case the impulse response of the electronic feedback filter needed to be long enough to model the acoustic feedback path. Any acoustic path in the enclosure has a resonant response; the impulse response at any position is representative of other positions so a digital controller able to model the enclosure response at one point would also be adequate to model the response at any other point in the enclosure. To model a system well the model needs to be long enough to encompass most of the energy of the system impulse response. Less and less payoff is to be had (in terms of fit against complexity) from extending the model beyond a certain accuracy.

The response of the feedforward filter approximates the inverse of a typical enclosure response (section 2.3). It is shown in section 2.3 that the inverse of a decaying second order system is shorter than the impulse response of the system itself. Therefore if the feedback filter is long enough to model an enclosure response then a feedforward filter of the same length should theoretically be long enough.

A loudspeaker at an enclosure corner was excited with a swept sine signal increasing in frequency from 0.1 to 500 Hz sampled at 1 KHz reconstructed via a low pass filter with a cut off frequency of 350 Hz. The response was captured at a microphone. The increase in length of the response over the input signal gave a measure of the typical impulse response length of the enclosure. The undamped response of the enclosure was of the order of 0.3 S.

To reduce the impulse response to the order of 0.12 S (short enough to be modelled with the 128 point filter sampling at 1 KHz) a small amount of foam (3 cm covering half of each face) was added to the ceiling and one end of the enclosure (figure 4.1.3). It was important not to damp the enclosure too much or the reverberant field that the control system was to work on would become unimportant compared with the direct field.

4.1.5 OPERATING LIMITATIONS OF THE CONTROLLER

The diagram of the control system (Figure 4.1.4) shows that the distance from the detector to the monitor is less than the distance from the control speaker to the monitor. This establishes the fact that this system could not attenuate the direct field over all frequencies but could only hope to cancel the reverberant field inside the enclosure.

The theory and practice applied here could equally well be applied to the attenuation of a travelling sound field provided the acoustic delay was greater than the group delay of the electronic control system.

It has been shown (Ref. Flockton among others) that with suitable technology plane waves travelling down a pipe can be successfully attenuated. Therefore this implementation was particularly concerned with the active control of the reverberant field in an enclosure; an area which had not been so well documented.

Figure 4.1.6 shows a list of sampling frequencies, appropriate cut off frequencies for the analogue filters and the corresponding group delay of the two filters used.

The total delay of the control system is estimated by adding the group delay, the processing delay and a millisecond estimate (typical order of magnitude) for the delay of the control speaker. Using a model of the sound propagating along a direct path from the monitor to the control speaker, an estimate is made of the minimum distance between the two speakers for the system to operate at this sampling frequency. The number of points needed in the digital filter is also given. For the direct field to be attenuated the physical dimensions, the sampling rate and the analogue delay through the control system need to give suitable conditions. When sampling at 1 kHz the microphone loudspeaker distance needed to be at least 1.7 m, much longer than the enclosure itself. The controller would need to run at over 10 KHz to have a hope of cancelling any of the direct field. This would require filter lengths of over 1000 points (more points are needed to cover the same time length when sampling at a higher frequency), much more than was available, so it was not practically possible to control the direct field inside the enclosure (these figures are based on the sound propagating in one direction, and this is not the case in an enclosed three dimensional space, but they serve as a guide).

4.1.6 POSITIONS OF THE TRANSDUCERS INSIDE THE ENCLOSURE

This demonstration was intended to give an insight into how an active control system could be implemented in a practical environment and to show that such a controller could be implemented successfully. The controller elements, the detector microphone and the control speaker were placed in "sensible" positions where theory and common sense indicated that the system could work.

A single detector, single source system was under test. A single source can successfully mimic the field from another single source. It cannot replicate the field created from two sources and using two or more speakers as a noise source would have introduced more degrees of freedom into the sound field resulting in a more complicated field to be controlled. The demonstration was concerned with testing and understanding a realisable control system; a single primary source was chosen and positioned in a corner so it would excite all the modes of the enclosure.

A single monitor position was chosen to avoid overcomplication. The microphone was placed in an arbitrary position roughly one third way along a diagonal from a corner so as to be sure of detecting the low order modes adequately. It was positioned in a quadrant of the box away from both speakers to be distant from the near field.

Following the rationale that any real field will be the sum of travelling and reverberant fields, the detector microphone is placed near the primary source. The nearer the detector is to the noise source the sooner the changes in both components of the sound field are detected.

The control speaker was positioned in another corner so as to be sure of being able to excite all modes within the working range of the control system.

The working frequency range of the controller had been designed to incorporate the first two modes of vibration of the enclosure; the 100 mode in the horizontal direction and the 010 mode in the vertical direction. Because the major concern of the control system was to attenuate these modes the enclosure can be thought of as a two dimensional system. The diagram of the control system in figure 4.1.4 is a plan looking down on the top of the box.

A picture of the enclosure and transducers is shown in figure 4.1.7. A projection of the side of the box is shown in figure 4.1.8. In this view the noise source and the control source become coincident; they are in the same position relative to the horizontal mode and the vertical mode. The view represents a vertical section through the enclosure; along any section the amplitude and phase of each mode is theoretically the same as that through any section. In practice, edge effects at the walls and coupling of the modes will alter the sound field but it is useful to ease the understanding of the sound field by regarding the field as two dimensional. In this respect the active control system is only operating on a two dimensional sound field. However, the difficult practicalities of implementing an active control system still make this a worthwhile demonstration.

4.1.7 SUMMARY

The acoustic system was conditioned and the sampling rate chosen in the steps described in order to design an experiment with suitable conditions in which an active noise controller could operate.

The hardware available determined the length of the digital filter that could be implemented. The reverberant field of the acoustic environment determined the desired working range of the system necessary to control the first two low order modes of the enclosure. The working range of the system determined the sampling frequency it needed to operate at. The sampling frequency and the available length of the digital filter determined the maximum length of impulse response that the system could model successfully using a finite impulse response filter. The acoustic environment was passively damped to limit the impulse response of the reverberant field to an appropriate length.

This final adjustment to the acoustic field would probably not have been possible in designing a system to actively control the field in a practical situation. A practical system would need to be designed in the following steps:

The frequency range of the noise levels to be attenuated would determine the required working range of the system which would determine the sampling frequency. The sampling frequency and the impulse response of the acoustic system would determine the hardware required.

Having described the initial conditions of the experiment the next section presents the practical derivation of the digital control filters and a theoretical assessment of their performance.

4.2 ACOUSTICAL MEASUREMENTS ON THE ENCLOSURE AND DERIVATION OF THE DIGITAL FILTERS

4.2.1 INTRODUCTION

This section describes the measurements and analysis performed on the test enclosure to determine the digital filters to be used in the controller. Measurements through various paths of the system are recorded and processed to derive the digital controllers. The digital controllers are compared to the theoretical spectra needed to give cancellation. A more precise understanding of the action of the controller is presented and the maximum theoretical attenuation achievable with the digital filters is derived.

4.2.2 GENERATION OF TRANSIENT TEST SIGNAL

The procedure used to derive the finite impulse response filters has been detailed in section 3.3. The method involved recording the response of various paths through the control system by exciting the system with a test signal and capturing the system response. The test signal (generated by the Fortran program in Appendix 4) consisted of a continuous sine wave rapidly increasing in frequency between two frequency limits in 0.128 S.

A signal was numerically generated linearly increasing in frequency range from 0.1 Hz to 500 Hz. As described in section 3.3 this range covers the intended working range of the control system and beyond the cut off frequency of the low pass filters up to the Nyquist frequency. The signal was comprised of 128 points at intervals of 1 mS. A signal which begins with a sharp onset and ends suddenly will have ripple in its spectrum due to the discontinuties in the signal; the amplitude spectrum can be flattened by increasing and decreasing the signal amplitude gradually at the start and end of the signal, so the first and last 20 points were modified by a cosine window.

The measurement system operated by outputting a signal of 512 points length and synchronously capturing an input signal of the same length. Therefore the total length of the signal was 512 points; the last 384 points of the numerically generated signal were of zero level leaving time for the captured response which would necessarily be longer (by a length equal to the impulse response of the system) after passing though the acoustic system.

The signal x_1 is shown with its corresponding spectrum X_1 in figure 4.2.3.

When working with time signals it is important to remember that because the signal length is finite its spectrum will not be correctly resolved (all frequency components will only be accurately resolved when a time signal of infinite length is transformed). Some aliasing will occur; this must be reduced to a manageable amount small enough to not significantly affect the analysis. Although it has been windowed the spectrum X1 is still significantly high at 500 Hz implying that frequency components just above the Nyquist frequency of 500 Hz are of the same magnitude; in the sampled time series these frequencies are incorrectly resolved as lower frequencies (aliasing). An easy way to reduce these high frequency components was to band limit the signal by passing it through an analogue low pass filter and recapturing the signal at exit from the filter. The signal x1 was passed through a low pass filter with a cut off frequency of 400 Hz. The captured signal, x_2 (figure 4.2.4) was used as the input signal for all subsequent measurements.

4.2.3 PROCEDURE TO RECORD ACOUSTIC MEASUREMENTS

The following procedure was used to record various measurements on the control system from which the digital controllers were derived. The apparatus was set up as in figure 4.2.1.

L0 is the noise source

L2 is the cancelling source

M1 is the detector microphone

M3 is the monitor microphone

The maximum input and output voltage levels to and from the TMS32020 microprocessor board were +/- 10 V. The output signal x_2 covered the complete dynamic range of the chip memory and hence resulted in a maximum output level of +/- 10 V at exit from the digital to analogue converter (DAC). The signal level at the analogue to digital converter (ADC) needed to be less than 10 V.

It was necessary that the signal levels in the system were conditioned so that the dynamic range of the microprocessor was not exceeded. The following procedure was used to set appropriate gain levels on the power amplifiers to the loudspeakers and the measuring amplifiers connected to the microphones;

By repeatedly driving the noise source L0 with the signal x_2, the resultant level at the detector microphone M1 was observed and the amplifiers A0 and A1 were adjusted to give a maximum signal level comfortably below 10 V modulus. Neither amplifier was set too low as to give a poor signal to noise level at that point. The acoustic path itself was probably the noisiest part of the system, especially as its response could change with for example a change in temperature and using high gain levels improves the S/N ratio through the acoustic path.

It was important that neither the amplifier gain levels nor the transducer positions were altered during measurements or implementation. The desired response of the controller was determined by the responses of the rest of the system and the derived filter responses were, within experimental constraints, the 'right answers' to that acoustic environment; adjusting a gain in the system would alter the controller needed to attenuate the field. This procedure fixed the amplifier A0 and A1 for the remainder of the demonstration.

The loudspeaker and microphone amplifiers were adjusted so that each speaker was of the same strength and each microphone was of the same strength. The sources were required to create the same pressure fields but in antiphase at the monitor position, consequently the amplifier A2 was adjusted so that source L2 gave the same pressure level at the monitor as when the field was excited by L0. The microphone gains were made approximately equal by placing both microphones in the same position, exciting one source and adjusting the amplifier A3. This procedure fixed the gains in the feedforward and feedback paths to be both of the order of unity (see section 3.2).

With the amplifier gains fixed the following measurements were recorded;

OUTPUT SIGNAL	x_2	x_2	y_{10}	x_2
SPEAKER	L0	L0	L2	L2
MICROPHONE	M1	M3	M3	M1
CAPTURED SIGNAL	y_{10}	y_{30}	y_{32}	y_1
FIGURE 4.2.	5	6	7	8

The following comments on the signals and their spectra are of interest. The graphics package used to display the signals (from Interactive Laboratory Systems inc.) could be used to take the Fourier transform of a signal and display the spectra with the signal alongside as shown. The length of each signal shown is 0.512 S, sampled at 1kHz. However, the signals themselves should be viewed with care. They are not displayed as representative heights but are scaled so the maximum amplitude covers the same vertical distance. The amplitude of the spectra, particularly the modal peaks can be used as a gauge of the signal amplitude. The dB scales on the spectra do not refer to SPL but are arbitrary levels which can only be compared to each other.

The modal nature of each spectrum is evident. Comparing the spectra Y_{10} and Y_{30} in the frequency range before the first mode shows higher levels for Y_{10}; at these frequencies the two low order modes are in phase at the detector position (this being in the same quadrant as the noise source), whereas at the monitor position the modes are out of phase, cancelling and resulting in the lower SPL at this position. This effect can also be seen at the dip in the spectra between the two modal resonances.

The speaker used for the noise source does not transmit frequencies well below about 100 Hz. The spectrum Y_{12} is similar to Y_{30} but is only well defined beyond 180 Hz. The control speaker resonated at a higher frequency than the noise source speaker.

The signal y_{32} is much longer than the others by the order of 0.1 to 0.2 S (the impulse response of the enclosure). This is because it is the result of passing through the acoustic system twice. The peaks in the spectrum are much more pronounced because the spectrum is the result of the broadband signal x_2 travelling twice through the acoustic bandpass system as opposed to the other signals passing through the system only once.

4.2.4 DERIVING AND ASSESSING THE DIGITAL CONTROL FILTERS

Each finite impulse response filter was derived by deconvolving two time series using a least squares approximation to obtain the best fit (see section 3.3). The feedback filter (figure 4.2.10) is the result of deconvolving signals x_2 and y_{12}. The feedforward filter (figure 4.2.12) is the result of deconvolving signals y_{32} and y_{30}. Each filter was 128 points long. The maximum coefficient magnitude in the feedback filter was 0.11 and the maximum magnitude in the feedforward filter was 1.56. It was shown in section 3.2 how, when operating in a reverberant field, the gain of one filter was theoretically inversely proportional to the gain of the other. However, the maximum coefficient values do not directly represent the gain in that path, especially as the feedforward filter is modelling a non-causal path.

The desired spectrum of the feedback filter is given by dividing Y_{12} by X_2. X_2 is flat over the working range so the ideal frequency response of the feedback filter (figure 4.2.11) is similar to Y_{12}. The amplitude response is dominated by the modal nature of the sound field. The phase response shows an overall linear characteristic from the acoustic delay and the group delay of the analogue filters. Superimposed on this are the rapid phase changes around the mode resonances; a phase change approaching a half cycle is noticable around 250 Hz at the first mode, as the response of the sound field, dominated by this mode, changes from leading the velocity of the speaker to lagging (section 2.1). A smaller phase change is noticeable around the second mode; but the shape of the spectra is evidently disrupted.

y_{12} was the result of passing x_2 through a filter (the acoustic system); y_{12} has resulted causally from x_2 (x_2 causes y_{12} to occur) and consequently the filter is totally causal (section 2.2). Therefore the derived digital filter closely resembles the inverse fourier transform of the spectrum Y_{12}/X_2 (cf. figures 4.2.10 and 4.2.11) the ideal impulse response needed to model the acoustic feedback path.

The digital filter needs to be long enough to contain most of the energy contained in the measured impulse response. It can be seen that this condition is satisfied; most of the energy in the impulse response is within the first 0.128 S.

Figure 4.2.9 shows the two spectra Y_{30} and Y_{32} superimposed. Over most of the frequency range the amplitude of Y_{32} lies below Y_{30} but goes above at the mode resonance frequencies. Y_{32} will have less energy than Y_{30} because it has passed through the filter system twice; but it will not necessarily have less energy at a particular frequency. The result of dividing Y_{30} by Y_{32} is shown in figure 4.2.13. Over the range of interest the amplitude response approximates a flat spectrum with dips at the modal frequencies as predicted in section 2.3.

Modelling the response between Y_{32} and Y_{30} ideally requires a partly non-causal filter; y_{32} does not cause y_{30} (section 2.2). Unlike the signals used to derive the feedback path, signal y_{30} is not a filtered version of signal y_{32}. The inverse transform of the spectra Y_{30}/Y_{32} (figure 4.2.13) does not give clear information about the digital filter required. What has happened ?

The acoustic path from the detector to the monitor has a shorter delay than the cancellation path so any filter implemented in the cancellation path will still result in the cancellation signal arriving too late to cancel the direct noise. What is needed is a filter which would cause the signal to jump ahead of time. This phenomenon is represented mathematically in the Fourier transform by extending the impulse response into negative time; ie. before the t = 0 axis, so the filtering mathematically occurs before it could actually happen with a real filter, which is what is required. Due to the circular nature of the functions calculated using a discrete Fourier transform this negative direction starts from the right hand side of the impulse response time series (figure 4.2.13) and continues backwards. As this part of the impulse response is superimposed on the causal part it is not possible to distinguish the two. If the causal and non causal parts of the response were shorter than the displayed length then it might be possible to distinguish them if they both decayed to a sufficiently low level before they overlapped. This can be useful when designing a filter to control a direct field; the proportion of energy contained in the non-causal part of the response represents the part of the response that cannot be practically realised and the extent to which the field cannot be attenuated.

Either part of the response may be longer than the time length of 0.512 S (determined by the resolution of the spectra; 1/0.512 = 1.95 Hz which in turn was determined by the length of the original signals). Because either, or both parts of the response are greater than 0.512 s then the time series gets 'wrapped around'; continuing at one end of the time axis where it finished at the other. Both parts of the impulse response appear to be very long and have wrapped round the axis a substantial distance. The spectrum of the filter shows two spikes around 125 Hz; these frequency components dominate the impulse response (which has a beating frequency of about 6 Hz; the spacing between the peaks). This illustrates the fact that a single spike in the frequency domain affects all of the time domain. This time series does not give useful information about the digital controller needed to attenuate the field.

The feedforward filter, derived by a least squared error fit of the time signal y_{32} to y_{30}, was a causal filter approximating a non-causal response. The filter and the ideal spectrum have similar shaped amplitude responses (figure 4.2.14) although they differ in magnitude considerably;- the causal filter is able to model the non-causal response with only a limited degree of success. A comparision of the phase spectra of the two responses is illuminating. It is known that to attenuate the direct field at the monitor position the filter needs to make the control signal jump ahead in time; this is represented by the overall trend of phase gain shown in the spectra of the ideal response. As shown, the causal finite impulse response filter cannot match this phase gain. Note the phase responses around 250 and 300 Hz; the mode resonances. Again the causal filter cannot match the non-causal system because it cannot cancel the direct field. However it can be seen that the phase response of the filter around the modes differs from the ideal response by approximately a half cycle. Two effects need to be considered here, the first one technical. The FIR filter had its coefficients negated to cause the control signal to be inverted and cancel the acoustic path. It could be said, therefore that the phase response of the filter matches the response of the non-causal filter around the acoustic resonances. However, this cannot be true, because as already stated the delay in the control path is greater than the delay in the acoustic path and the phase responses cannot match.

The signal from the detector contributes to a particular cycle of the resonance of the first mode. This same signal from the detector is processed and generates, in antiphase, a cycle of the same mode; but this cycle will occur an odd number of half cycles later than that due to the original signal. It is the highly resonant nature of the mode, approximating a sinusoidal response that enables a half cycle from the control source to attenuate a half cycle from the noise source even though it is arriving a number of half cycles later; it attenuates the mode by cancelling with a different half cycle from the noise source.

4.2.5 THEORETICAL ATTENUATION ACHIEVABLE ON THE REVERBERANT FIELD

It has been discussed how, because the acoustic feedback path has a causal response it can be adequately modelled if a long enough filter is used. The feedforward filter, which is modelling a non-causal response is the unknown quantity here. A theoretical simulation was carried out to determine the maximum attenuation that could be achieved with this feedforward filter. The feedback filter was assumed to work perfectly (ie. the feedback paths were ignored). Figure 4.2.15 shows the spectrum Y_{30} captured at the monitor microphone when driving the noise source with a swept sine signal. This spectrum can be regarded as the spectrum at the monitor when the noise source is driven with a broadband signal. Superimposed on this is a plot of the theoretical spectrum at the monitor position with the controller on.

The attenuated spectrum was calculated as follows with the notation used in sections 2.2 and 3.1.

The spectrum at the monitor microphone is given by;

$$P_2 = P_{o2} + P_1 \, T \, C$$

In this analysis the spectrum at the monitor due to the noise source alone (P_{o2}) can be represented by Y_{30}. The spectrum at the detector (P_1) can be represented by Y_{10}. Ignoring the feedback, the controller T is given by the response of the feedforward filter FF. The response of the path from the control speaker to the monitor (C) is given by dividing the measured spectra Y_{32}/Y_{10} . Hence the spectrum at the monitor is given by;

$$P_2 = Y_{30} + FF \, Y_{32}$$

The spectrum at the monitor was calculated by adding the spectrun Y_{30} to the product of the spectrum Y_{32} with that of the feedforward filter.

The theoretical spectrum at the monitor with the controller on shows the first two low order modes significantly attenuated by about 10 dB each at the resonant frequencies. The remainder of the spectrum is left largely unaffected; the controller cannot cancel the direct field, but importantly, does not theoretically enhance the direct field at this position. This 128 point filter was used in the practical implementation in section 4.3; it can be seen how the theoretical response compares well with the actual response in figure 4.3.2. Figure 4.2.16 shows a 256 point filter derived from signal y_{32} and y_{30}. The theoretical maximum attenuation given by this filter is shown in figure 4.2.17. The increase in filter length has not significantly increased the attenuation; a 128 point filter is long enough to model the causal part of the response and produce the maximum attenuation achievable in these physical circumstances.

The high frequency region of figure 4.2.17 is not a valid prediction of the spectra with or without the controller because the measuerments were band limited with the low pass analogue filters. The simulation does not predict the effect of the controller at high frequencies outside the working range. It can be seen (figure 4.2.12) that the feedforward filter has a high amplitude response at high frequencies. It is thought that this has resulted from the FIR identification program modelling a non-causal response. The actual system response at high frequencies will only be correctly specified if the low pass filters attenuate this region to a significant extent.

At frequencies between the modal peaks the sound field is dominated by two modes ie. the field has two degrees of freedom. It is not possible to control the field at these frequencies with a one degree of freedom controller and as the modes were already cancelling the best solution at this monitor position would have been to leave these frequencies unfiltered.

It was shown in the last section how a phase delay exists between the two modal frequencies in the FIR filter whereas the phase response of the non-causal theoretical response shows a phase lead. Although the cancellation signal can be in phase at the modal frequencies because of the opposite phase trends then the cancellation signal will necessarily be out of phase with the non-causal ideal and attenuation will not be achieved over this frequency range.

However this is the controller that gives the best fit from this model; it is suggested that because the mode resonances dominate the signals y_{32} and y_{30} the FIR modelling routine gives most weight to modelling the response at these frequencies; rightly so. Consider the action of the controller at a particular frequency; at frequencies around the first resonance the controller is primarily concerned with attenuating the first mode, because it is this mode that dominates the sound field, a similar situation occurs at frequencies around the second resonance. In both cases the detector lies in the same quarter wavelength as the control source (see figures 4.2.18 and 19) and so (ignoring the responses of the transducers and the electrical system) the filter needs to delay the signal by an odd number of half cycles to attenuate each mode seperately. The optimum solution using the least squared error model was to delay the frequencies dominating the first mode by a certain number of half cycles and the frequencies dominating the second mode by an additional two half cycles. Given this solution, at frequencies inbetween the mode resonances the controller needs to delay a frequency by so many half cycles to attenuate one mode and an additional cycle to work on the other mode; it cannot do both. This highlights a previously unforseen difficulty in using the least squared error routine to derive the digital filters.

Summarising the above, there seem to be two reasons for the controller not attenuating the pressure level over a frequency range inbetween the two modal frequencies. Firstly, complete cancellation cannot be achieved because the theoretical controller needed is non-causal. Secondly, because of the poor signal to noise ratio arithmetic inaccuracies result in the FIR modelling being insufficient at these frequencies. Doubling the filter length still does not prevent the enhancement at the inter modal frequencies. This suggests that it is the non-causality; the underlying physical problem that causes this enhancement. The one degree of freedom control system cannot control the frequency range where the field is dominated by two degrees of freedom.

4.2.6 SUMMARY

The swept sine signal used as the input signal for the acoustical measurements was generated and suitably conditioned. A procedure was developed to successfully record the measurements on the enclosure. The digital control filters were derived by a least squared error fit; it was shown how suitable lengths had been chosen for the filters and their form was explained by theory.

The results have given a more precise understanding of the action of the controller in particular how two low order modes can be attenuated. It is shown how, theoretically the single detector, single source controller will attenuate the first two modes of the reverberant field in the enclosure.

The digital controller was derived and its theoretical performance evaluated; the next section presents some experimental measurements recording the actual attenuation when the control system was practically implemented.

4.3 PRACTICAL IMPLEMENTATION OF THE ACTIVE CONTROL SYSTEM

4.3.1 INTRODUCTION

Having obtained filter coefficient values for the digital controllers and also predicted their performance in the previous section, this section presents some practical measurements used to assess the degree of success of the control system.

Transfer function measurements of the closed loop and open loop system are presented to assess the stability of the controller. The results of the practical implementation of a single detector, single source control system operating in the enclosure are presented. The attenuation given by the system at the monitor position is shown as well as the effect of the controller at another position.

4.3.2 CLOSED LOOP TRANSFER FUNCTION MEASUREMENTS ON THE CONTROL SYSTEM

It is instructive to quantify the degree of stability of a control system and whilst implementing the working system various measurements were taken to try to assess the degree and areas of instability of the controller. Assessing the stability of the controller involves assessing the stability over the whole frequency range as the degree of instability is determined by the least stable frequency. Although the measurements of closed loop responses presented here do not give a quantifiable measure of the system stability they give a strong indication of the match of the digital feedback filter to the acoustic feedback.

The results of three separate experiments measuring various closed loop responses of the system are presented. The first measures the match between the amplitude spectra of the two feedback filters. The second verifies the match of the phase responses and the third demonstrates the successful match of the paths with the feedforward filter present in the system.

A primary requirement of the controller is that it remains stable. The stability is initially governed by the match of the feedback paths; if the electronic path perfectly cancelled the acoustic feedback path then no matter what filter was in the forward path then the system would remain stable. An investigation was made into the cancellation achieved by the digital feedback path using the apparatus shown in figure 4.3.1. The noise source loudspeaker was still present in the enclosure but only the elements of concern here are shown in the diagram. The spectrum analyser was used to inject noise into the system before the digital controller and sample the resultant signal at exit from the controller. Various closed loop transfer functions were recorded.

The concept of how the feedback controller is to operate is shown in figure 4.3.2; with both feedback paths operating the system is designed to be stable. However, with only one feedback path in operation stability is not ensured; indeed this returns to the original situation of needing to overcome the instability introduced by the acoustic feedback. The experimental arrangement did not quite correspond to this block diagram. The actual arrangement is shown in figure 4.3.3. The necessary difference is introduced because the signal needed to be band limited at entrance to and exit from the digital system.

The first coefficient in the digital feedforward path was made non-zero and the remaining taps set to zero to make the filter an all pass filter up to the cut off frequency of the low pass filters. The transfer function was recorded (figure 4.3.4).

Both feedback paths were now connected to the system; ie. the acoustic path connected and the digital coefficients loaded. The resulting closed loop transfer function of the system in figure 4.3.1 is shown in figure 4.3.5. The similarities in the amplitude spectra are evident indicating the successful cancellation of the feedback paths. The phase spectra differ because the responses are measured over different paths. The difference is due to the fact that the transfer function was recorded at a different point from where the feedback paths met and cancelled.

The limitations on the digital filter matching the acoustic path are;

1. The digital filter is of a finite length and is inevitably truncated with respect to the impulse response of the acoustic path. This was probably the most serious limitation on the match of the digital system.

2. The control system was implemented and the closed loop measurements taken at a later time from when the measurements were recorded to derive the digital filters; the characteristics of the system will change slightly over time.

3. The digital filter is a numerical model of the real filter, in this case a set of coefficients resulting in a signal with the least squared error over time compared to the signal from the acoustic path; these coefficients themselves are subject to rounding errors as they are implemented as integers. The overall resolution of the digital system was determined by the resolution of the analogue to digital and digital to analogue converters. Although the microprocessor operated with 16 bit arithmetic the ADC and DAC resolution was only 12 bit

determining a signal to noise ratio of less than 78 dB. This is the quantisation signal to noise ratio resulting from the analogue signals being quantised to specific digital values. It was estimated that the sample and hold devices connected to the ADC and DAC had a S/N ratio of the order of 50 dB.

4. Both low pass filters used to band limit the digital system had eight order roll offs, equivalent to 96dB/octave. Using filters with cut off frequencies at 350 Hz the S/N ratio would be about 40 dB by 500 Hz. Any acoustic or electrical noise in the system with a SN ratio less than about 40 dB would have affected the system accuracy.

A second closed loop experiment is presented to verify the phase response match of the digital feedback filter. Measurements were recorded of the transfer functions of the seperate feedback paths shown in figures 4.3.6 and 4.3.7. The responses are shown in figures 4.3.8 and 4.3.9. The responses of the two feedback paths are different because the measurements are of two different systems. Let the following transfer functions be represented by the following notation;

H1 feedforward digital path

H2 feedback digital path

H3 acoustic feedback path

H the two analogue filters

The transfer function of figure 4.3.6 is given by;

$$\frac{H\ H1}{1 - H1\ H2}$$

The transfer function of figure 4.3.7 is given by;

$$\frac{H\ H1}{1 - H\ H1\ H3}$$

The digital feedback path is designed so that it models the response of the acoustic feedback path and the analogue filters, but the signal in the digital system is also inverted; ie. multiplied by minus one. Hence the modelling the digital path is trying to achieve is given by;

$$H2 = - H\ H3$$

Therefore the responses of figures 4.3.6 and 4.3.7 given above are not equal. The analysis shows that if the digital feedback path or the acoustic feedback path is inverted then the responses should be equal. Practically the easiest way to do this was to reverse the terminals on the control loudspeaker. The response of this transfer function is shown in figure 4.3.10. The match with figure 4.3.8 is evident thereby supporting the analysis presented above verifying the match of the phase spectra and again verifying the match of the amplitude of the feedback filters.

The feedforward filter was loaded into the forward path and the closed loop responses with and without both feedback paths connected were recorded (figures 4.3.11 & 4.3.12). The shape of the feedforward filter was as expected (cf. figure 4.2.12) and again connecting the feedback paths did not significantly affect the response. In this case the response could not be recorded with only one feedback path connected because the loop was unstable.

All these measurements indicate the success of the feedback filter in cancelling the acoustic feedback. However the measurements do not give a quantitative measure of the stability of the system. The way to measure this would have been to implement the system and increase the gain in the forward path until the system became unstable. However, the dynamic range of the digital system did not make this possible. Therefore a program was written (see section 3.3 and appendix 6) to run on the microprocessor to measure the open loop transfer function round a loop.

4.3.3 OPEN LOOP TRANSFER FUNCTION MEASUREMENTS ON THE CONTROL SYSTEM

Measurements of the open loop transfer function round a loop were recorded to gain an insight into the degree of instability of the controller and at what frequencies it was most likely to go unstable. The results of three experiments are presented. The first predicts the open loop response of the system with a single non-zero coefficient present in the feedforward filter. The second experiment measures this response. The third experiment measures the open loop response with the feedforward filter in the system.

The swept sine signal (described in section 4.2) was injected into the system at a point immediately before the forward filter and the resultant signal captured at the same position (figure 4.3.13).

A single coefficient of unity was placed in the forward path and the acoustic path disconnected to measure the response of the implemented feedback path only (figure 4.3.14). The resultant signal was denoted yd_{12} (figure 4.3.15), the digital counterpart of y_{12}. The signals y_{12} and yd_{12} were added and the resultant signal is shown in figure 4.3.16; a significant reduction in the height of the modal peaks can be seen. The spectrum of this signal was divided by the spectrum of the swept sine signal resulting in the open loop transfer function through the system (figure 4.3.17).

A system will be unstable when the magnitude of the loop gain is greater than one and the phase shift round the loop is an odd number of half cycles. It can be seen from figure 4.3.17 that the phase shift round the loop is very variable and for a half cycle shift to occur round the loop is quite common. Therefore the chance of the system slipping into instability is essentially determined by whether or not the gain round the loop ever exceeds one. If this is the case then it would be quite easy for the phase change round the loop to vary to such an extent as to give instability. The diagram shows that the system is liable to become unstable at very low frequencies below about 20 Hz. Over the rest of the spectrum the gain round the loop is less than -20 dB (ie 0.1).

The digital feedback path was derived from the swept sine signal and the signal y_{12}. It was expected that the feedback paths show a good match. It is noticable how the mode resonances are not completely cancelled, this can be attributed to the finite length of the filter which is truncated with respect to the acoustic system, the resonances being the part of the impulse response that rings on longest. The relatively high levels at low frequencies can be seen in the shape of the digital feedback filter itself (figure 4.2.10). The levels are probably due to this as the loudspeaker does not transmit low frequency components well. The poor modelling of the feedback path at these low frequencies can be attributed to the fact that the swept sine spectrum falls off at the low frequencies. However this spectrum has also been used to generate the open loop transfer function itself so the accuracy of the function at these low frequencies is questionable. A solution to this would have been to use a transient signal with a defined spectra at these low frequencies. Alternatively, if low frequency matching of the feedback paths was a problem then using a feedforward filter with low gain at low frequencies would compensate for this. The problem remains of how to generate such a filter.

A measurement was recorded of the response round the system with both feedback paths connected (figure 4.3.18). The captured signal is shown in figure 4.3.19 and the resulting open loop transfer function in figure 4.3.20. The noisiness of the spectrum compared to the theoretical match is evident, particularly around the mode resonances where the response is 10 dB worse. The system is more likely to be unstable at very low frequencies.

Interpretation of the signals themselves should be made with care as each signal is scaled to the same amplitude to be displayed: the very messy signal of figure 4.3.19 is mostly noise containing the resonant components within it.

The open loop transfer function of the complete system, with feedforward filter included was recorded (figures 4.3.21 & 4.3.22). Again the response shows decreased stability at low frequency. While setting up the working system it became obvious that low frequency instability was causing difficulty. Initially measurements on the acoustic system had been recorded using a swept sine signal ranging from 10 Hz to 500 Hz as opposed to a signal starting at 0.1 Hz in the successful implementation. The controller showed a similar feedback match (figure 4.3.23 cf. 4.3.20) but the open loop function of the complete system (forward filter included) shows a higher gain round the loop at low frequencies (figure 4.3.24 cf. 4.3.22). Indeed, this controller was unstable. When the controller was implemented it could be viewed on an oscilloscope how the output from the digital system would be saturated for a time then it would decrease to a realistic level when a single spectrum on the analyser would show that the reverberant field had been attenuated at that instant. Saturation would then occur again, driving the control speaker very hard and enhancing the noise levels. This pattern was repeated with a time period of seconds indicating that the instability was occuring at very low frequencies. This showed that the system was made stable by conducting the initial measurements on the system with a swept sine signal of frequency range down to 0.1 Hz.

4.3.4 RESULTING ATTENUATION AT THE MONITOR POSITION WHEN THE CONTROL SYSTEM WAS IMPLEMENTED

The single detector, single source control system is shown in figure 4.3.25. The noise source was driven with a pseudo random signal from a Hewlett Packard HP3582A spectrum analyser and the signal from the monitor microphone connected to the analyser to record the transfer function (averaged 256 times) between the signal driving the noise source and the signal at the monitor. The amplitude of the response at the monitor microphone with and without the controller is shown in figure 4.3.26.

The amplitude of the controlled noise levels agrees well with the theoretical predictions of figure 4.2.15; significant attenuation occuring only at the mode resonances. Levels outside the working range of the controller are left largely unaffected; the low pass filters at the entrance to and exit from the digital system have removed the high frequency components in the signal sufficiently.

The relative phase change between the noise source and the signal at the monitor is shown with and without the controller in figure 4.3.27. It can be seen how the sharp phase changes at the modes are removed by the action of the controller: because a large part of the modal nature of the field has been cancelled at these frequencies the phase changes due to the resonances have been removed.

The control system was stable and attenuated the field to the same extent months after the system had been set up and the digital filters had been derived: demonstrating substantial stability over time.

The effect of the control system at the monitor position (figures 4.3.26 and 27) and at another observation position (figures 4.3.30 and 31) is shown. The attenuation of the reverberant field is investigated, however no attempt was made to determine the overall reduction in the space averaged sound pressure level inside the enclosure.

4.3.5 EFFECT OF THE CONTROLLER AT ANOTHER POSITION IN THE ENCLOSURE

The frequency components around the first two modes of the enclosure were attenuated at the monitor position. These frequencies excite the first two modes of the enclosure and as they had been removed at one position then theoretically the modes themselves had been attenuated, ie. the standing waves were no longer present in the enclosure.

To illustrate this phenomenon the field was viewed at another position in the enclosure. The observation position was chosen to be in a different quadrant of the enclosure from the monitor and detector positions. Diagrams and pictures are shown in figures 4.3.28 and 4.3.29. The noise level with and without the controller is shown in figure 4.3.30 (and the phase spectra in figure 4.3.31). It can be seen that the frequencies around the mode resonances have been attenuated. This demonstrates the success of the controller in removing the reverberant field in the enclosure; as predicted attenuating these frequencies at one position attenuates them throughout the enclosure because these frequencies make up the standing wave pattern inside the enclosure.

However, at the observation position the noise levels at frequencies inbetween the mode resonances have been enhanced. The resultant noise level at the monitor position was explained by considering the action of the controller in section 4.2. Over the frequency range of a single mode which is well defined in frequency relative to other modes the acoustic system can be approximated to a single degree of freedom system which can be controlled by a single controller. However at frequencies inbetween the mode resonances the acoustic system is dominated by two vibrations; ie. has two degrees of freedom; such a system cannot be adequately controlled by a controller with only one degree of freedom.

The important fact demonstrated here is that the first two modes of the reverberant field had been significantly attenuated and this attenuation has been shown to occur at two representative points in the enclosure, not just at the point that the system was designed to operate at.

4.3.6 SUMMARY

The closed loop transfer function measurements indicated the success of the digital system in modelling the controller required. However they did not give a quantifiable measure of the degree of success of the control system. Therefore the open loop transfer function was measured to give a measure of the degree of stability.

Finally, the actual attenuation achieved when the control system was implemented was presented.

The implementation of the single detector, single source control system successfully attenuated the first two resonances of the reverberant field of the enclosure. However, the single degree of freedom controller could not control the field at frequencies where the field was dominated by two modes, ie. had two degrees of freedom.

4.4 ANALYSIS OF THE CONTROL SYSTEM AT DIFFERING SAMPLING FREQUENCIES AND MONITOR POSITIONS. AND IMPLEMENTATION OF A TWO CHANNEL ACTIVE CONTROL SYSTEM.

4.4.1 INTRODUCTION

This section presents some additional experiments performed on the single channel control system operating inside the enclosure and the implementation of a two channel active control system consisting of a single detector and two speakers.

Measurements were taken at different sampling rates with different degrees of damping present in the enclosure and the theoretical attenuation achievable from a single channel control system under such conditions derived. This analysis practically demonstrates the performance limitations of the active control system.

The implementation of a single channel system is presented (as in the experiment forming the bulk of this chapter) but designed to give attenuation at a different position to that in section 4.3.

A series of measurements is presented by which the control filters for a single detector, double source system can be derived. These measurements were performed experimentally and the results are discussed.

This section refers to results and discussions presented in previous sections, particularly sections 4.2 and 4.3. For this reason it would be difficult to read the bulk of this section out of this context.

4.4.2 ANALYSIS OF THE CONTROLLER PERFORMANCE AT DIFFERENT SAMPLING RATES AND DEGREES OF DAMPING

It was demonstrated in section 4.1 how it was not possible using the available signal processing hardware to control the direct field in this enclosure given the small acoustic travel time across the enclosure compared to the group delay of the analogue filters and the loudspeaker. Some experiments were performed to verify this and observe if the controller had a significantly detrimental effect on the field at the monitor.

The enclosure was heavily damped by lagging all the walls and the ceiling with 6 cm of foam. This removed a large proportion of the reverberant field leaving the direct field dominant in the enclosure. A series of measurements at 1, 5 and 10 KHz are shown in figures 4.4.1, 2 and 3. Each series of measurements shows the signals (and spectra) y_{10}, y_{30}, y_{32}, the ratio of the spectra Y_{30}/Y_{32} and the impulse response of this spectrum (derived from the inverse fourier transform). Lastly, the

128 point FIR feedforward filter derived from the signals y_{32} and y_{30} is shown.

The reduced signal lengths and flatness of the spctra are evident, although resonances are still apparent. Comparisons of the ideal non-causal spectra and causal model show the similarities which are discussed in the main experiment of section 4.3. Plots of the theoretical attenuation achievable (ignoring the feedback) are shown in figure 4.4.4. These plots show that no worthwhile attenuation is achieved but the response at the monitor microphone is not significantly enhanced.

4.4.3 IMPLEMENTATION OF A SINGLE CHANNEL SYSTEM CONTROLLING THE FIELD AT A DIFFERENT MONITOR POSITION

The experiment presented in section 4.3 described the implementation of a single detector, single speaker control system attenuating the field at a particular point inside the enclosure. It was shown in section 3.1 how moving the monitor microphone does not alter the acoustic feedback path. Therefore the digital feedback path does not need to be altered, only the feedforward filter need be adjusted.

An experiment was performed where the roles of the monitor and observation microphones of section 4.3 were reversed. The digital controller was redesigned to control the sound field at the new monitor position. The feedback filter remains the same. However this experiment was carried out at a different time and all measurements were repeated and both digital filters redesigned. The signal measurements are shown in figure 4.4.5.

The signals y_{10} and y_{12} compare with the corresponding signals in the experiment of section 4.2 and 4.3 (cf. figures 4.2.5 & 4.2.8). Signals y_{30} and y_{32} differ from figures 4.2.6 and 4.2.7 because of the different position of the monitor.

The 128 point finite impulse response filters and their responses are shown in figure 4.4.6. The non-causal spectra resulting from dividing Y_{30} by Y_{32} is also shown. The feedback filter is similar to that of figure 4.2.10 and the feedforward filter differs from figure 4.2.12.

The amplitude and phase responses at the monitor position with and without the controller are shown in figure 4.4.7.

There is a noticable difference between the controlled response at the monitor and the response at the previous monitor position (figure 4.3.30). In the previous case the best solution for frequencies inbetween the modal peaks was to leave those frequencies unfiltered (see section 4.2); the effect of the controller in the previous case was to enhance those frequencies.

It is thought that the increased signal levels in the spectra Y_{30} and Y_{32} have resulted in the least squared error modelling routine giving a good fit for the feedforward filter at these frequencies. At the new monitor position the modes are in phase at frequencies inbetween the mode resonances and the controller does not have a detrimental effect.

The good match of the feedforward filter at the frequencies inbetween the resonances is also demonstrated in the response at the observation microphone (previously the monitor) shown in figure 4.4.8. The modal frequencies have been attenuated as always and, in addition, inbetween the resonances the signal level has remained largely unchanged.

However significant enhancement is seen at frequencies immediately before the first mode. A similar effect has occured as at the inter modal frequencies in the first experiment; without the controller the two significant modes were cancelling over these frequencies and the controller, being unable to control the two degrees of freedom has resulted in the favourable cancellation being destroyed. To be sure of being able to control the two degree of freedom sound field it is necessary to use a two degree of freedom control system.

4.4.4 PRACTICAL ARRANGEMENT OF THE TWO CHANNEL CONTROL SYSTEM

It was shown in section 3.1 how the digital controller for a single channel control system could be used as the basic building block to implement a general multichannel control system.

Having successfully implemented a single channel control system experimental work proceeded to demonstrate the implementation of a two channel control system operating on the sound field inside the small enclosure.

The control system consisted of a single detector microphone and two control loudspeakers positioned as in figure 4.4.9, the side projection of figure 4.4.10 and the picture of figure 4.3.29.

As discussed in section 4.1, because the noise source is a single loudspeaker, it is appropriate to use a single detector microphone. In this case however because two control speakers were to be used it as appropriate to monitor the field at two positions. If only one monitor were used then the digital controller needed to operate the system would no longer be unique; it is indeterminate. The control filter feeding one of the speakers can be arbitrarily chosen and the other then chosen to give the desired result at the monitor. Practically it is not realistic to implement such a system.

It was shown in section 2.3 how the monitor positions need to be appropriately chosen so that they independently monitor the field and all the dominant modes are monitored. The two monitor positions used in practice were the same two positions used in the main experiment (the single monitor and observation position). The analysis in figure 2.3.10 shows two monitor positions which give an indeterminate solution. A similar analysis in figure 4.4.10 shows that the determinant of the characteristic functions of the monitor positions is non-zero implying that the positions were independent.

Having described the practical configuration of the control system it is necessary to show how the digital filters for the two channel control system were derived. The method is similar in essence to the series of measurements presented in section 3.3 and in the practical implementation of section 4.2. Therefore the next subsection is of a theoretical nature and can be left unread without detracting from the practical information presented in this chapter.

4.4.5 SERIES OF MEASUREMENTS USED TO DERIVE THE DIGITAL CONTROLLERS FOR THE TWO CHANNEL CONTROL SYSTEM

Figure 4.4.11 is used to illustrate the control system; the diagram can be used to aid the visualisation of the measurements undertaken.

The feedback filters can be simply measured by exciting each control speaker in turn and capturing the response at the detector. The pairs of signals then need to be deconvolved to derive the filters.

The feedforward filters are given by (section 3.1);

$$K = C^{-1} A B^{-1}$$

The matrix equation can be expanded to give;

$$k_1 = \frac{a_1 c_{22} - a_2 c_{12}}{B\,(c_{11} c_{22} - c_{12} c_{21})}$$

$$k_2 = \frac{a_2 c_{11} - a_1 c_{21}}{B\,(c_{11} c_{22} - c_{12} c_{21})}$$

where k_1 and k_2 are the forward filters to the first and second control speakers. It can be seen that the denominator of both these terms is the same; this term is the equivalent of the signal y_{32} in the derivation of the single channel controller. The numerator of the equations (equivalent to the the y_{30} signal) differs.

The method available for deriving the control filters was that of deconvolving two time series. Therefore the result of any measurements on the control system needed to be two time series. A summary of the measurements performed is presented below. The captured signals are listed together with the input signal, loudspeaker and microphone used for the measurements and the transfer function recorded.

CAPTURED SIGNAL	INPUT SIGNAL	LOUDSPEAKER	MICROPHONE	TRANSFER FUNCTION
ya_{12}	x_2	LA	D	f_1
yb_{12}	x_2	LB	D	f_2
y_{10}	x_2	LO	D	B
a_{30}	x_2	LO	MA	a_1
b_{30}	x_2	LO	MB	a_2
a_{32}	y_{10}	LA	MA	$B\ c_{11}$
b_{32}	y_{10}	LA	MB	$B\ c_{21}$
ba_{32}	a_{32}	LB	MB	$B\ c_{11}\ c_{22}$
ab_{32}	b_{32}	LB	MA	$B\ c_{21}\ c_{12}$
aa_{30}	a_{30}	LA	MB	$a_1\ c_{21}$
ab_{30}	b_{30}	LA	MA	$a_2\ c_{11}$
ba_{30}	a_{30}	LB	MB	$a_1\ c_{22}$
bb_{30}	b_{30}	LB	MA	$a_2\ c_{12}$

where c_{21} denotes the transfer function from the first loudspeaker to the second monitor (notation of section 3.1).

The three signals that are needed to derive the feedforward filters can be derived from;

$ya_{30} = ab_{30} - aa_{30}$

$yb_{30} = ba_{30} - bb_{30}$

$y_{32} = ba_{32} - ab_{32}$

These measurements were recorded on the control system described above and the two feedforward and two feedback filters (from ya_{12}, yb_{12} and x_2) derived by deconvolving the time series.

4.4.6 RESULT OF THE PRACTICAL IMPLEMENTATION OF THE TWO CHANNEL CONTROL SYSTEM

The implementation of the two channel control system resulted in a stable system. This in itself was a success; it showed that the two interacting feedback paths were decoupled and stabilised by the controller configuration (figure 4.4.9).

However, the control system did not attenuate the sound levels at the monitor positions; the enhanced sound levels are shown in figure 4.4.12.

Some additional analysis helps to explain what has possibly happened. Consider the configuration of transducers shown in figure 4.4.10. It is demonstrated in the figure how, using representative values for the positions of the microphones;

$$\begin{pmatrix} P1 \\ P2 \end{pmatrix} = \begin{pmatrix} -0.7 & -0.7 \\ -0.7 & 0.7 \end{pmatrix} \begin{pmatrix} A1 \\ A2 \end{pmatrix}$$

However, the mode amplitudes A1 and A2 are excited by the loudspeakers LO, LA and LB at positions 0,0 0,0 and X,0. Let the speakers have strengths QO, QA and QB.

The excitation of two modes from two sources is considered first before the more complicated sound field from three sources. Consider how the mode amplitudes are excited using only loudspeaker LA (as in the experiment forming the bulk of this chapter) and LO;

$$\begin{pmatrix} A1 \\ A2 \end{pmatrix} = \begin{pmatrix} 1 & 1 \\ 1 & 1 \end{pmatrix} \begin{pmatrix} QA \\ QO \end{pmatrix}$$

This equation demonstrates that the pressure at the monitor positions due to the control speaker is the same as the pressure due to the noise source loudspeaker. This is evident because the speakers are coincident. Therefore it is easy to use this speaker as the control source by making the source strengths have opposite signs.

Now consider how the mode amplitudes are excited by the second control loudspeaker and LO;

$$\begin{pmatrix} A1 \\ A2 \end{pmatrix} = \begin{pmatrix} -1 & 1 \\ 1 & 1 \end{pmatrix} \begin{pmatrix} QB \\ QO \end{pmatrix}$$

Hence;

$$\begin{pmatrix} P1 \\ P2 \end{pmatrix} = \begin{pmatrix} 0 & 1.4 \\ 1.4 & 0 \end{pmatrix} \begin{pmatrix} QB \\ QO \end{pmatrix}$$

The first equation demonstrates how the two speakers driven with the same source strength drive the first mode of amplitude A1, in antiphase. This is evident because the speakers are at opposite ends of the 1,0,0 mode. If the analysis is performed considering the modes seperately then it can be shown how the elements of the matrix are comprised of a term from each mode such that;

$$\begin{pmatrix} P1 \\ P2 \end{pmatrix} = \begin{pmatrix} 0.7 + -0.7 , & 0.7 + 0.7 \\ 0.7 + 0.7 , & -0.7 + 0.7 \end{pmatrix} \begin{pmatrix} QB \\ QO \end{pmatrix}$$

It may be considered that around the modal frequencies only one mode is dominant and only one term of the matrix in the above equation will apply. However, the implication of this equation is unclear; it serves to demonstrate the more complicated theory representing the two speaker control system where;

$$\begin{pmatrix} A1 \\ A2 \end{pmatrix} = \begin{pmatrix} 1 & -1 & 1 \\ 1 & 1 & 1 \end{pmatrix} \begin{pmatrix} QA \\ QB \\ QO \end{pmatrix}$$

resulting in;

$$\begin{pmatrix} P1 \\ P2 \end{pmatrix} = \begin{pmatrix} -1.4 & 0 & -1.4 \\ 0 & -1.4 & 0 \end{pmatrix} \begin{pmatrix} QA \\ QB \\ QO \end{pmatrix}$$

It is thought that the experiment was ill-conditioned and did not result in a control system producing attenuation because the second control loudspeaker was positioned in an inappropriate place. Consider the analogous case of two microphones monitoring almost the same sound; the situation will not be indeterminate but the two simultaneous equations will be very similar and noise will greatly affect the validity of the result. In reality the zero terms in the matrix above will be small non-zero values (the monitor positions were not exactly equal to the numerical values used) and what may have happened was that one of the forward filters was derived from a small signal resulting from the difference of two sizeable signals.

Theoretically it is advantageous to control the two modes with two control speakers. However the system did not succeed. The sound field resulting from each loudspeaker can be thought of as resulting from that source and an infinite number of images. The sound field produced by three sources will be the result of three such interference patterns. Perhaps the position of the speakers was inappropriate or the modes sufficiently distinct in frequency such that using two control speakers has overcomplicated the system. A good test of the two channel control system would be to implement the system operating on the sound field from two independent noise sources; this would require two detector microphones and there was not sufficient processing time available to implement a system with eight control filters on the TMS32020 microprocessor.

The situation is very complicated with a number of factors (transducer positions and number of modes within the system working range) contributing to the resultant sound field; the experimental work was concluded having reached a plateau of understanding sufficient to show how the low order modes can be attenuated in a simple experimental situation and demonstrating how the modal pattern greatly complicates any practical situation.

4.4.7 SUMMARY

It has been demonstrated how the control system cannot operate outside certain limits of the sampling frequency and damping present in the enclosure. The system does not attenuate the direct sound field inside the enclosure. The reproducability of the method of implementing the single detector, single speaker control system is verified and the conclusions of sections 4.2 and 4.3 supported.

At frequencies where the sound field is dominated by two modes the single degree of freedom controller may enhance the sound pressure level depending on the position at which the controller is designed to control the field.

5. CONCLUSIONS

5.1 REVIEW OF THE THEORY PRESENTED IN THIS THESIS

Active control systems can only operate by working to attenuate the sound field at one or more sensor locations. In chapter 2 it was shown how, to globally attenuate a sound field or produce a volume of attenuation it is necessary that these points be representative of the sound field throughout the volume of interest. In the case of a reverberant field it is necessary that the microphones pick up sufficient information about the dominant modes. Section 2.3 considered the requirements of an active sound control system operating on a reverberant field inside an enclosure. It was shown how n modes can be monitored with n monitor microphones and a mathematical treatment was used to demonstrate how these monitor positions need to be independent. It was also shown how n modes can in theory be controlled with a control system consisting of n channels each comprising of a single detector microphone and a single control speaker, each channel coupling into at least one mode of the sound field.

The active attenuation of a single mode was studied in section 2.3; the controller operation was studied in the frequency and time domains and the degree of attenuation achievable by a digital system of specific length considered.

Section 3.1 presented the theory giving the desired responses of the controllers for an active control system consisting of a number of detectors and speakers controlling the field at a number of monitor positions. The controller response was derived for the simplist version of such a control system consisting of a single detector and a single speaker capable of controlling the field at a single monitor position. It was shown how the controller could consist of a pair of electronic filters; one between the detector and the source in parallel with another cancelling the acoustic feedback from the detector.

The responses of some specfic multichannel controllers were derived and a suitable configuration of filters derived, such that any multichannel controller can be readily realised.

It was shown how the complexity of the controllers is reduced if the number of monitors is the same as the number of sources. It was also shown how increasing the number of or moving the monitor microphones does not alter the acoustic feedback present in the system. Therefore designing a controller with independent feedback compensation infers that the monitors can be adjusted without having to adjust the feedback compensation.

The major result of section 3.1 was to show a simple topology for multichannel controllers. The method relies on summing the digital feedback paths at the start of the digital system before the point where the signal divides to enter the feedforward filters prior to the control speakers. The acoustic feedback paths are thereby directly cancelled by the electronic feedback paths, which themselves are assured of being causal filters. The method demonstrates how any multichannel controller can be readily realised by repeatedly using a number of the same type of filter pairs used in the single channel system; thereby gaining considerable advantages for the practical development (hardware) of multichannel systems. However, it was noted that the number of filter pairs is equal to the square of the number of channels, thus limiting the number that could be implemented in practice.

5.2 REVIEW OF THE DIGITAL SYSTEMS IMPLEMENTED.

The same hardware was used both to record the various frequency responses of the system used to derive the digital filters for the controller and also to implement the digital filters. Additional hardware was developed to enable two channels to be input and output simultaneously.

In section 3.2 it was shown how, to ensure stability, it is necessary that the digital feedback filter gives a good match to the acoustic feedback over the entire working frequency range and also that outside this range the feedback gain is sufficiently low. It was shown how this could be achieved by implementing both the feedback and the feedforward filters on a single digital system. The practical aspects of the implementation of this system were presented, particular attention being paid to the operation of the convolutions within the controller and the testing of the program listed in appendix 2. It was verified that the measurement procedure and the filter implementation were compatible in their manipulation of the time series and filter coefficients and that the procedure resulted in successful system modelling.

The suitablity of the measurement method was discussed in section 3.3. It was observed that it was necessary to define the frequency response over the entire working range of the system; ie. beyond the cut-off frequency of the low pass filters and how an easy way to enable the electronic control system to compensate for its own operation is to record the acoustical system responses with the same hardware system used to implement the control system. It was also discussed how it was desirable to operate on data in the form of a time series.

The practical method used to derive the filter coefficients for a single channel active control system was presented in section 3.3. The method consisted of a series of acoustical measurements on the control system. The measurements were performed by exciting the control system with a swept frequency sine wave output from the digital system and capturing the response on the same system (program of Appendix 5) and deconvolving pairs of signals using a least squared error fit to derive the FIR filters. The open loop transfer function round a single channel control system was also measured and a digital controller to operate a single input, double output control system was implemented.

5.3 DISCUSSION OF THE RESULTS OF CHAPTER FOUR.

The main results of the practical work presented in this thesis were contained within chapter four, the first three sections of which were concerned with the implementation of a broadband active sound control system operating in a small enclosure. Section 1 was concerned with conditioning an experimental situation in which an active control system could be made to work with the available hardware. The loudspeakers used had dimensions much less than the half wavelengths of the modes to be attenuated and could therefore, to a first approximation be regarded as point sources. The practical implementation represented a simple case of the theory presented in section 2.3 and the successful implementation thereby supported this theory.

Section 4.2 presented the measurements and analysis performed to derive the digital control filters and the filters and their operation were explained. The theoretical attenuation produced by 128 point and 256 point feedforward filters was derived. It was shown how the longer filter did not significantly improve the attenuation and it was infered from this that the filter was adaquately modelling the causal part of the spectrum and the non-causality of the system was the fundamental limit to the attenuation achievable.

Section 4.3 presented various measurements used to assess the degree of stability of the control system governed by the match of digital and acoustic feedback paths. Closed loop transfer function measurements indicated the match of these paths and open loop transfer function measurements indicated the stability of the control system. The measurements and the practical implementation of the control system demonstrated how the test signal used to derive the filter coefficients needed to be defined over the entire working range of the control system.

The control system consisting of a single detector microphone and a single control speaker successfully attenuated the first two low order modes of vibration of the sound field inside the enclosure.

The demonstration showed how a controller can be successfully implemented on a single microprocessor; it was shown in section 3.1 how this single channel controller may be regarded as the basic unit of a multichannel controller. A control system consisting of a single detector and two control sources did not attenuate the sound field . However, the system was stable thereby demonstrating the success of the digital feedback compensation used.

The single channel control system could not control the sound field at frequencies where it was dominated by two modes of vibration. This was explained in terms of the frequency response of the feedforward filter in section

4.2. This agrees with the conclusions of Nelson (Ref).

The implemention demonstrated that the single channel control system did not significantly enhance the direct field inside the enclosure. This was also demonstrated by measurements and analysis performed at higher sampling frequencies in section 4.4.

5.4 GENERAL CONCLUSIONS AND SUGGESTIONS FOR FURTHER WORK

The successful implementation of an active sound control system, over a narrow frequency range in a simple test enclosure demonstrates how such a system may be implemented over a broader frequency range in a more practical enclosure to attenuate the reverberant field.

A practical system requires the control of the sound field at a number of monitor positions representative of the sound field. To understand how such a system may be implemented requires further study of the operation of a two channel controller. Also a study of adaptive digital filtering can indicate how the control system can be made adaptive.

FIGURES

$$P_{(X,Y,Z,W,T)} = \sum_{l}\sum_{m}\sum_{n} \underline{\varepsilon_x\varepsilon_y\varepsilon_z \frac{W c^2}{2\ Wn\ Kn + j(W^2 - Wn^2)}}\ \underline{\sum \frac{\rho\ Q_{(X0\ Y0\ Z0)}}{V} e^{j(wt+\phi)} \cos(l\pi\ X0/a) \cos(m\pi\ Y0/b) \cos(n\pi\ Z0/d)}\ \underline{\cos(l\pi\ x/a) \cos(m\pi\ y/b) \cos(n\pi\ z/d)}$$

Pressure at time t at a point x,y, vibrating at angular frequency w

Pressure is the sum of pressure components from each mode

COMPLEX MODE FUNCTION determining the frequency response of the mode

COMPLEX SOURCE STRENGTH DISTRIBUTION

combine to produce COMPLEX MODE AMPLITUDE

CHARACTERISTIC FUNCTION OF THE MODE

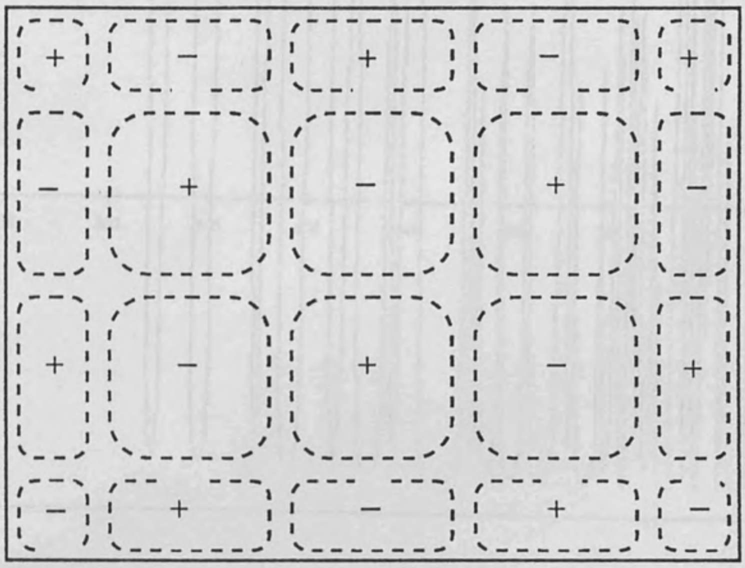
$\varepsilon_x\varepsilon_y\varepsilon_z$ DIRECTIONAL COUPLING COEFFICIENTS DETERMINED BY THE PARTICULAR MODE EACH COEFFICIENT DEPENDS ON THE MODE ORDER AND IS EQUAL TO 2 UNLESS THE MODE ORDER IN THAT DIRECTION IS 0 THEN THE COEFFICIENT IS EQUAL TO 1

Wn	MODAL FREQUENCY $= \Pi\ c\ [(l/a)^2 + (m/b)^2 + (n/d)^2]^{0.5}$
V	VOLUME OF ENCLOSURE = a b d (x,y & z dimensions)
Kn	DAMPING FACTOR (= 0 FOR AN UNDAMPED MODE)
X0,Y0,Z0	POSITION OF SOUND SOURCE
Q	VOLUME VELOCITY OF SOURCE
C	VELOCITY OF SOUND

TO DETERMINE THE PRESSURE AT A POINT THE MODAL PRESSURES AT THAT POINT ARE SUMMED OVER ALL THE MODES OF THE ENCLOSURE. l,m and n are the mode orders in the x,y and z directions. THEREFORE ALL THE COMPONENTS WITHIN THE MAIN SUMMATION SIGNS REFER TO QUANTITIES SPECIFIC TO A PARTICULAR MODE.

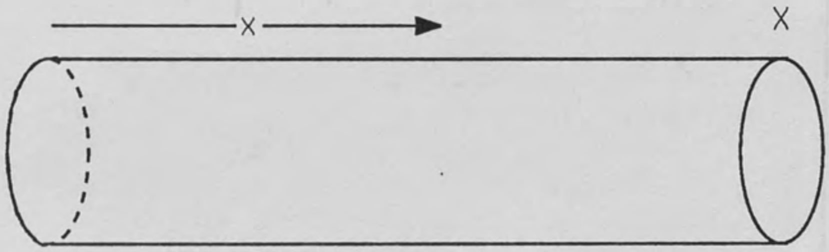
THE PRESSURE AT A POINT INSIDE A RECTANGULAR ENCLOSURE

FIGURE 2.1.1



A 2 DIMENSIONAL STANDING WAVE PATTERN
SHOWING RELATIVE PHASES

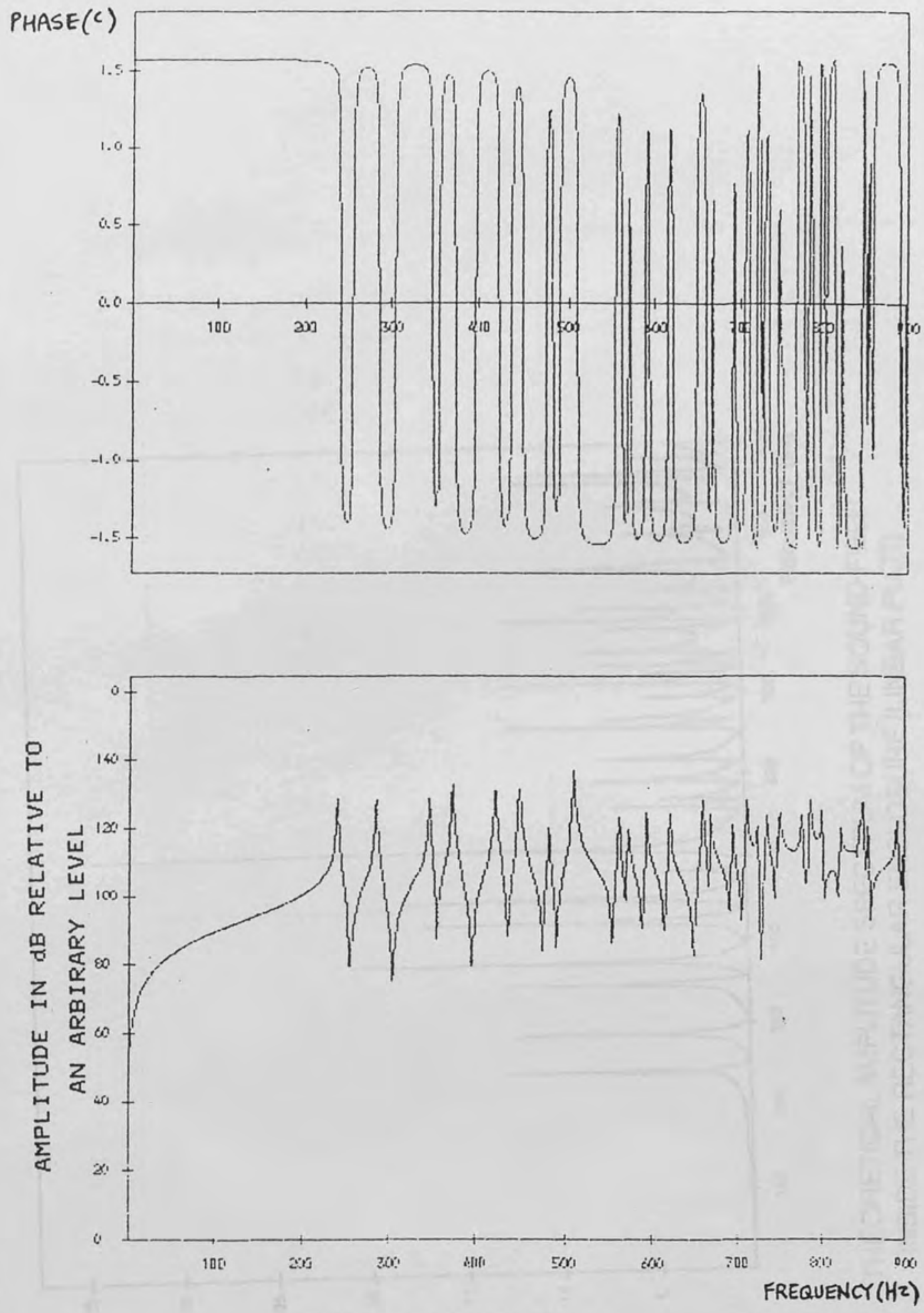
FIGURE 2.1.2



$$P = A \cos(2\pi x/X)\, e^{jwt}$$

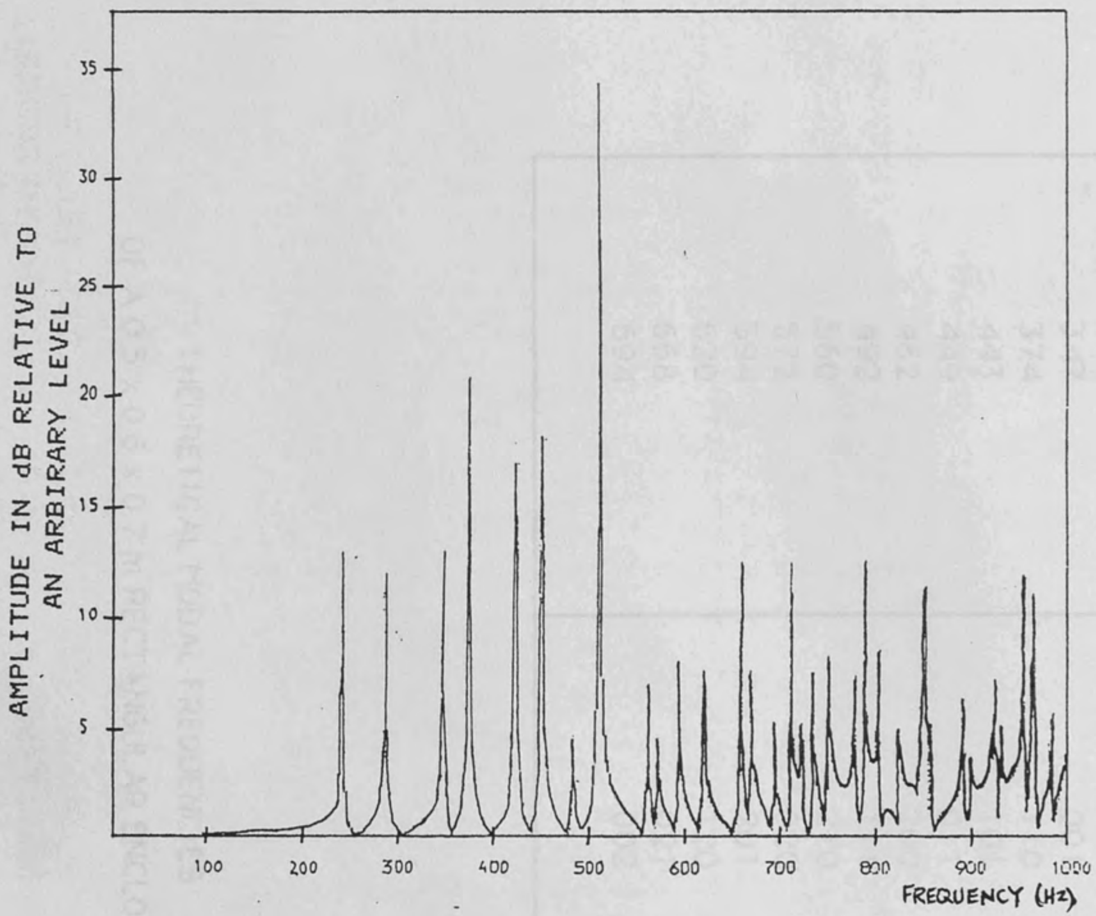
STANDING WAVE IN A CLOSED PIPE

FIGURE 2.1.3



THEORETICAL AMPLITUDE AND PHASE SPECTRA
OF THE SOUND FIELD
INSIDE THE RECTANGULAR ENCLOSURE

FIGURE 2.1.4.



THEORETICAL AMPLITUDE SPECTRUM OF THE SOUND FIELD
INSIDE THE RECTANGULAR ENCLOSURE (LINEAR PLOT)

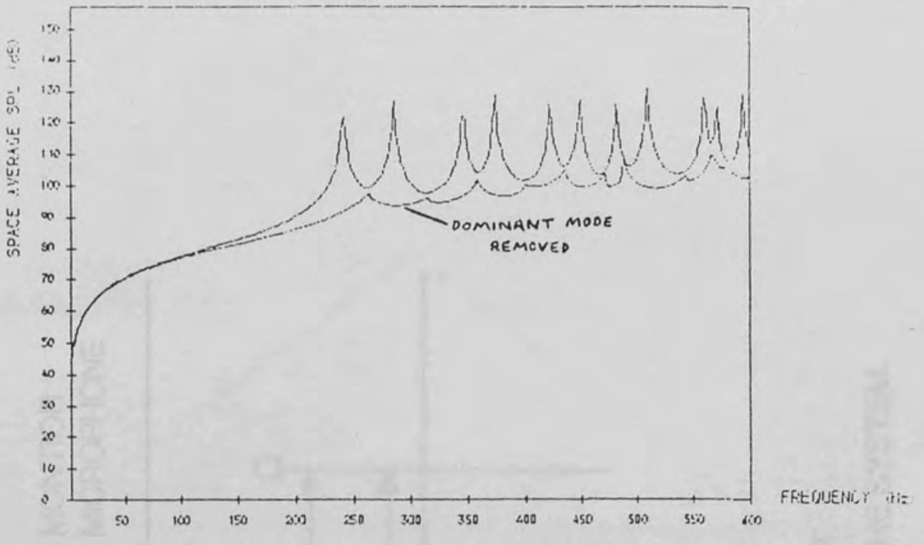
FIGURE 2.1.5

Frequency (Hz)	mode order
241	100
286	010
347	001
374	110
443	101
449	011
482	200
492	111
560	210
572	020
594	201
620	120
668	021
694	002

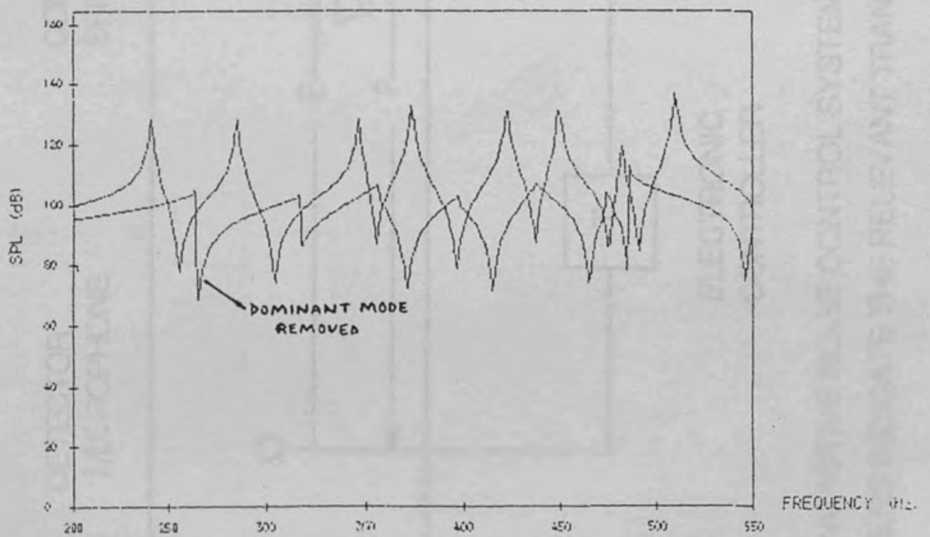
THEORETICAL MODAL FREQUENCIES

OF A 0.5 x 0.6 x 0.7 m RECTANGULAR ENCLOSURE

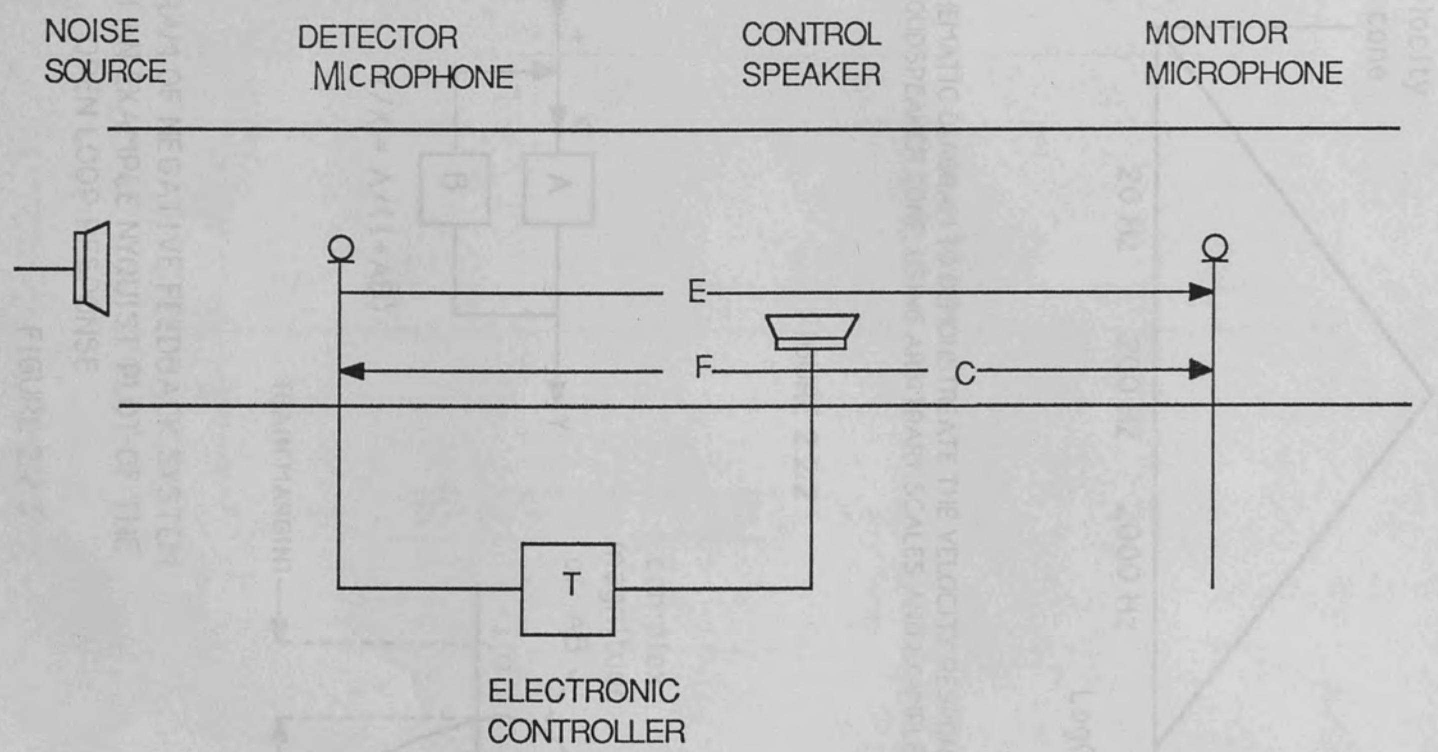
FIGURE 2.1.6



THEORETICAL AMPLITUDE SPECTRUM
OF THE SPACE AVERAGED SOUND PRESSURE LEVEL IN THE ENCLOSURE;
SHOWING THE EFFECT OF REMOVING THE LOCALLY DOMINANT MODE
AT ALL FREQUENCIES
FIGURE 2.1.7



THEORETICAL AMPLITUDE SPECTRUM
OF THE SOUND PRESSURE LEVEL AT A POINT IN THE ENCLOSURE;
SHOWING THE EFFECT OF REMOVING THE LOCALLY DOMINANT MODE
AT ALL FREQUENCIES
FIGURE 2.1.8



AN ACTIVE NOISE CONTROL SYSTEM OPERATING IN A DUCT.

THE LETTERS INDICATE THE RELEVANT TRANSFER FUNCTIONS OF THE SYSTEM.

FIGURE 2.2.1

SCHEMATIC DIAGRAM TO DEMONSTREATE THE VELOCITY RESPONSE OF
A LOUDSPEAKER CONE; USING ARBITRARY SCALES AND EXAMPLE VALUES

FIGURE 2.2.2

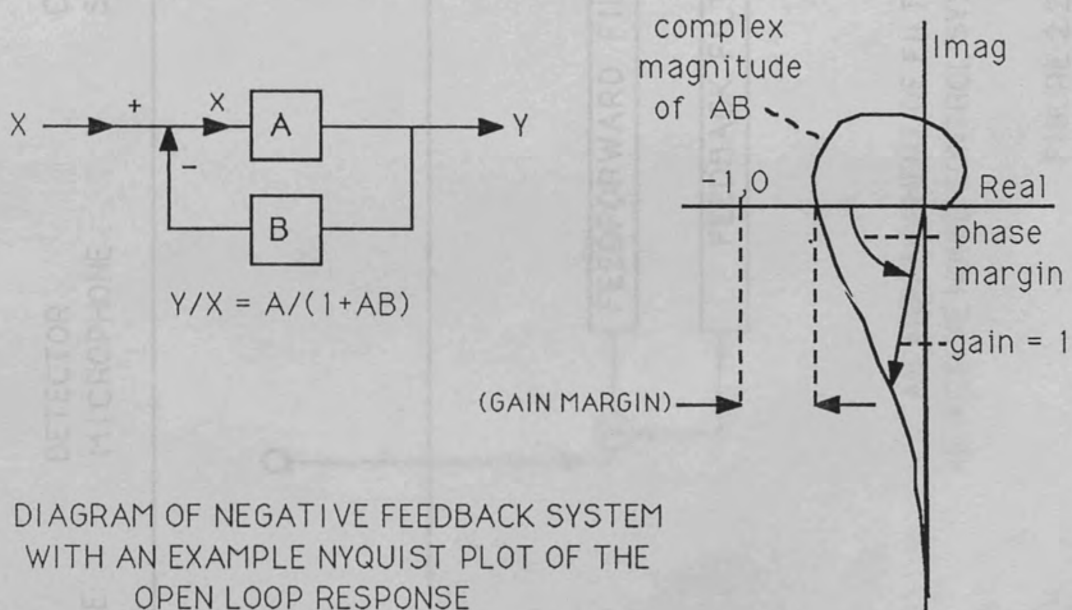
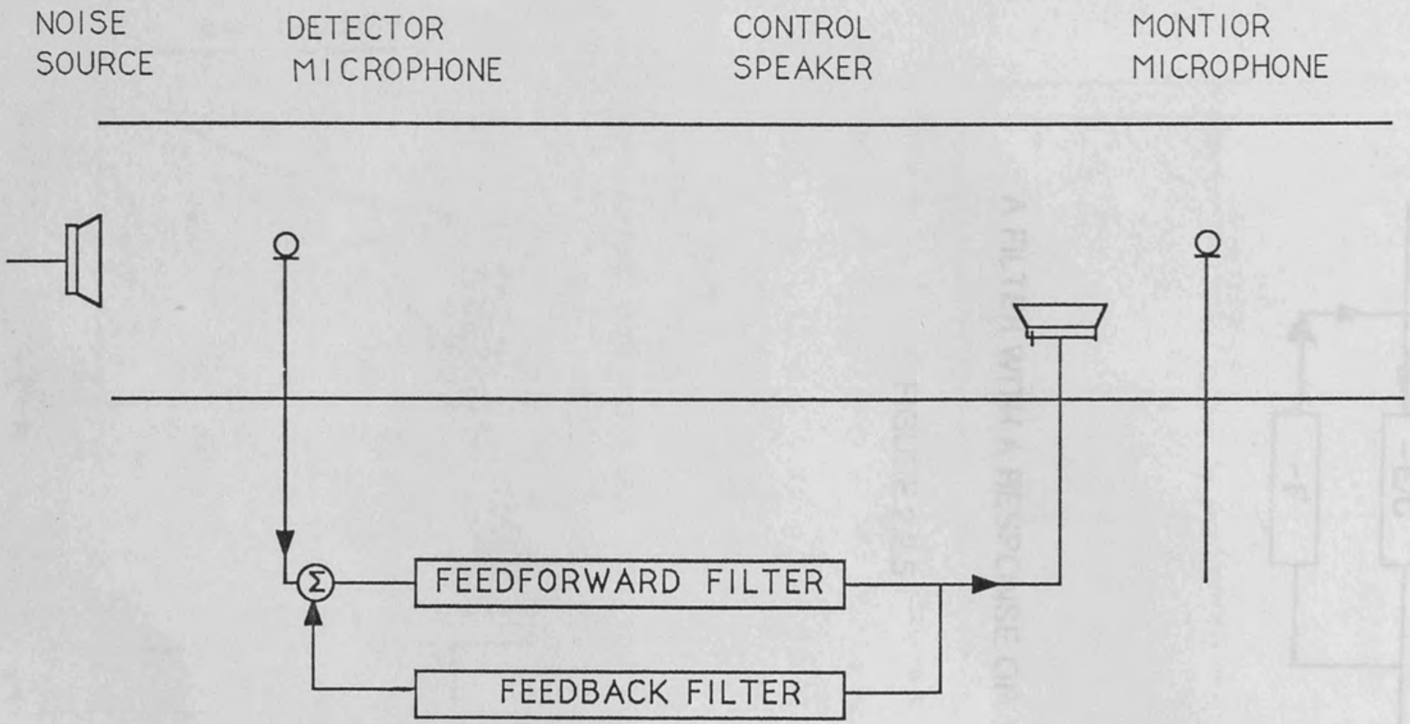


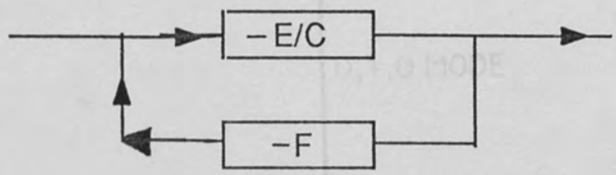
DIAGRAM OF NEGATIVE FEEDBACK SYSTEM
WITH AN EXAMPLE NYQUIST PLOT OF THE
OPEN LOOP RESPONSE

FIGURE 2.2.3



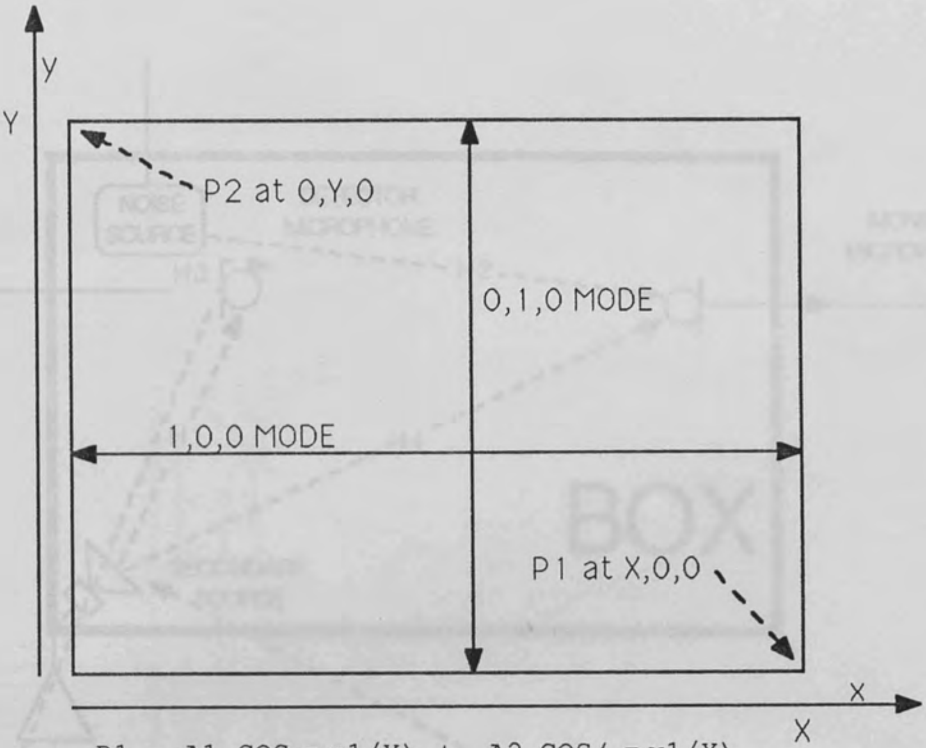
AN ARRANGEMENT OF FILTERS USED TO CONTROL
AN ACTIVE NOISE CONTROL SYSTEM OPERATING IN A DUCT.

FIGURE 2.2.4



A FILTER WITH A RESPONSE OF 1/(F-C/E)

FIGURE 2.2.5



$$P1 = A1\ COS(\pi\, x1/X) + A2\ COS(\pi\, y1/Y)$$

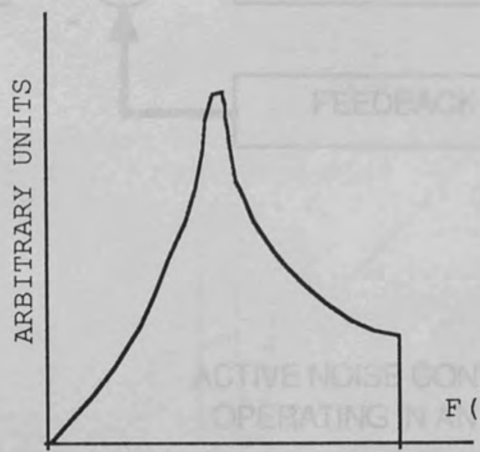
$$P2 = A1\ COS(\pi\, x2/X) + A2\ COS(\pi\, y2/Y)$$

$$X1 = X \quad X2 = 0 \quad Y1 = 0 \quad Y2 = Y$$

$$\mathbf{P} = \Psi\ \mathbf{A} \qquad \Psi = \begin{pmatrix} -1 & 1 \\ 1 & -1 \end{pmatrix}$$

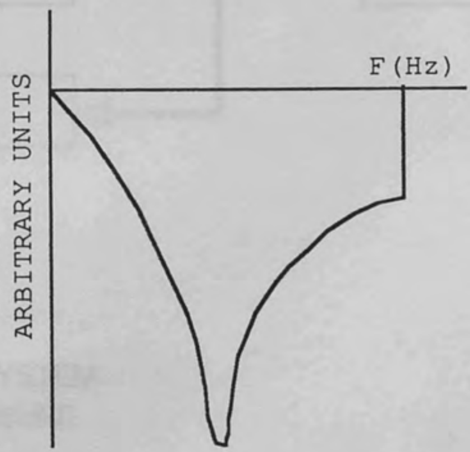
DEMONSTRATING INAPPROPRIATE MONITOR POSITIONS

FIGURE 2.3.1



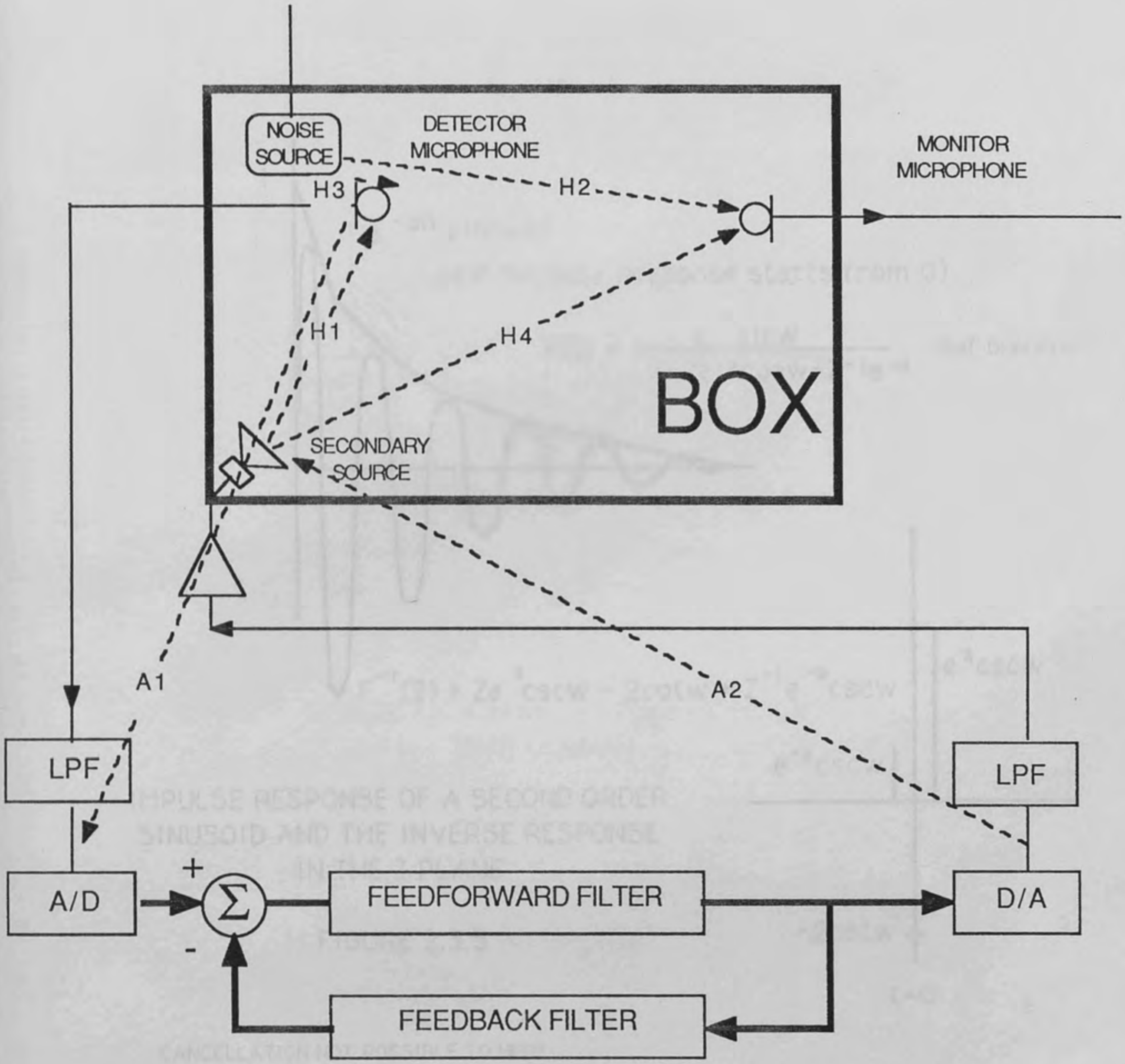
AMPLITUDE RESPONSE
OF A MODE

FIGURE 2.3.2



AMPLITUDE RESPONSE
OF THE FEEDFORWARD
FILTER TO CANCEL
A MODE

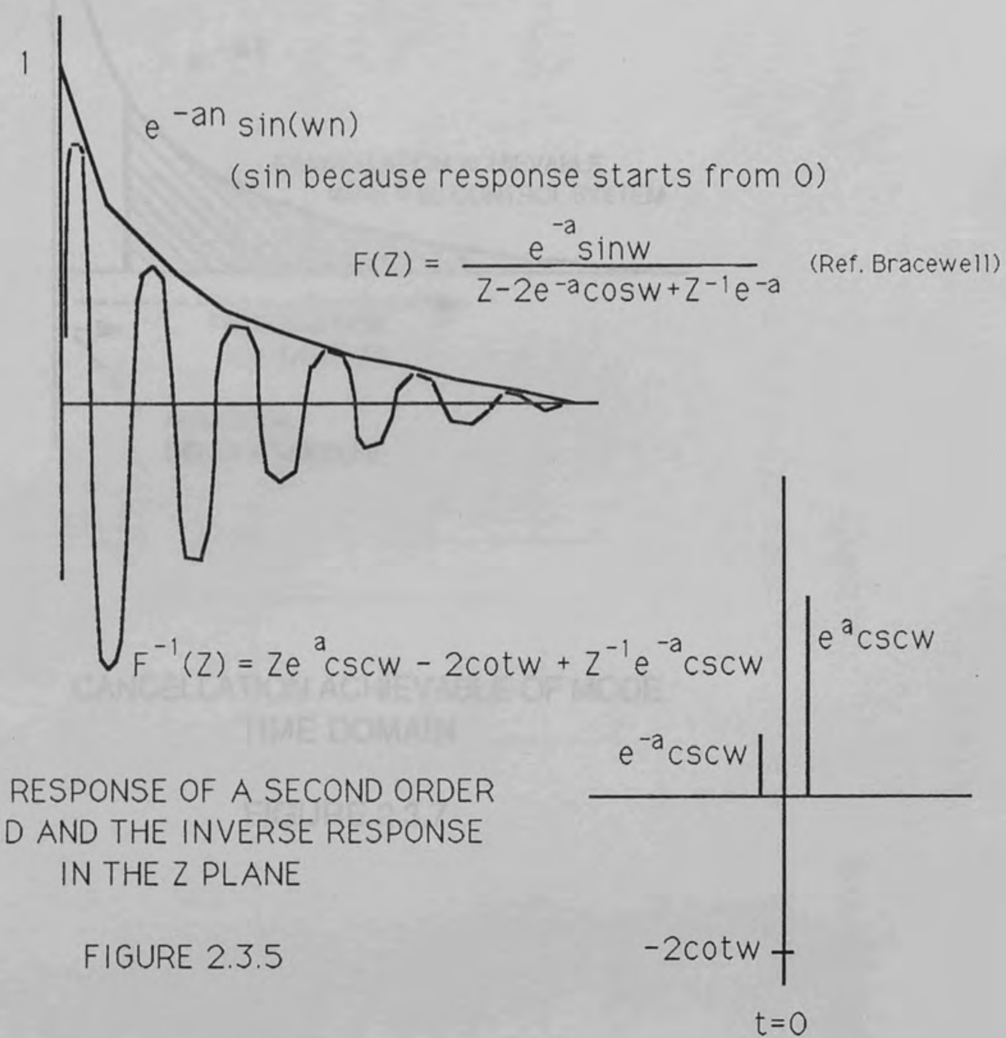
FIGURE 2.3.4



ACTIVE NOISE CONTROL SYSTEM
OPERATING IN AN ENCLOSURE

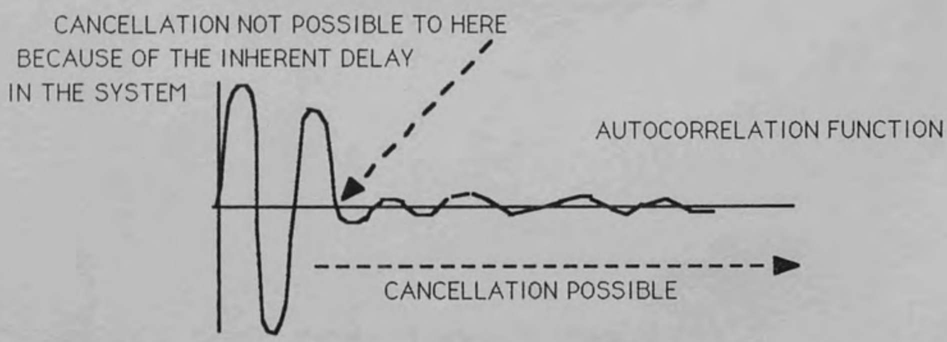
SHOWING RELEVANT TRANSFER FUNCTIONS

FIGURE 2.3.3



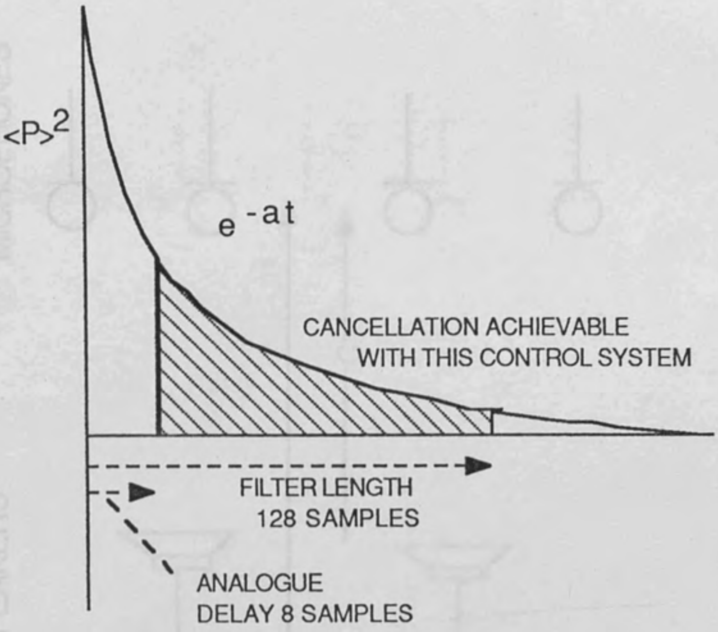
IMPULSE RESPONSE OF A SECOND ORDER SINUSOID AND THE INVERSE RESPONSE IN THE Z PLANE

FIGURE 2.3.5



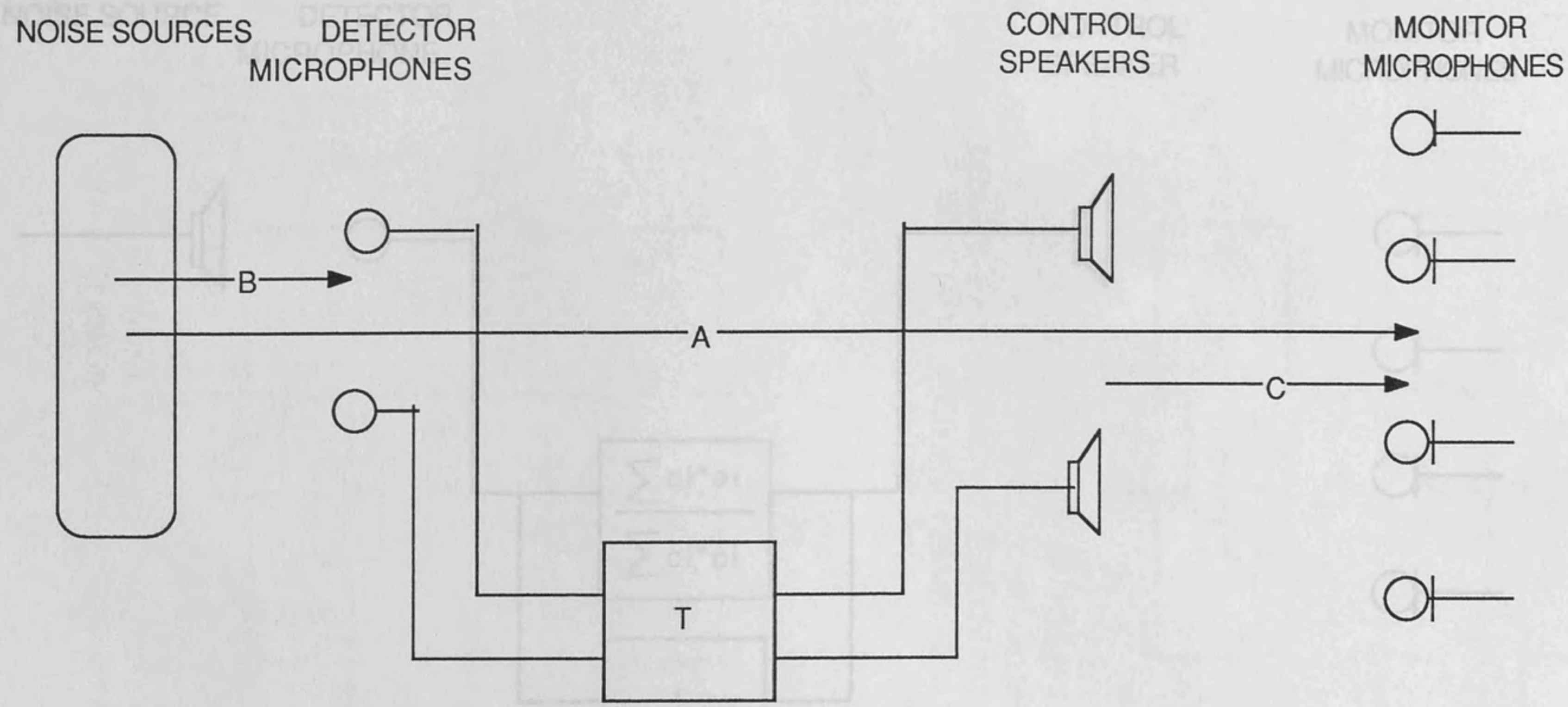
THEORETICAL ATTENUATION OF THE TIME RESPONSE OF A MODE

FIGURE 2.3.6



CANCELLATION ACHIEVABLE OF MODE
TIME DOMAIN

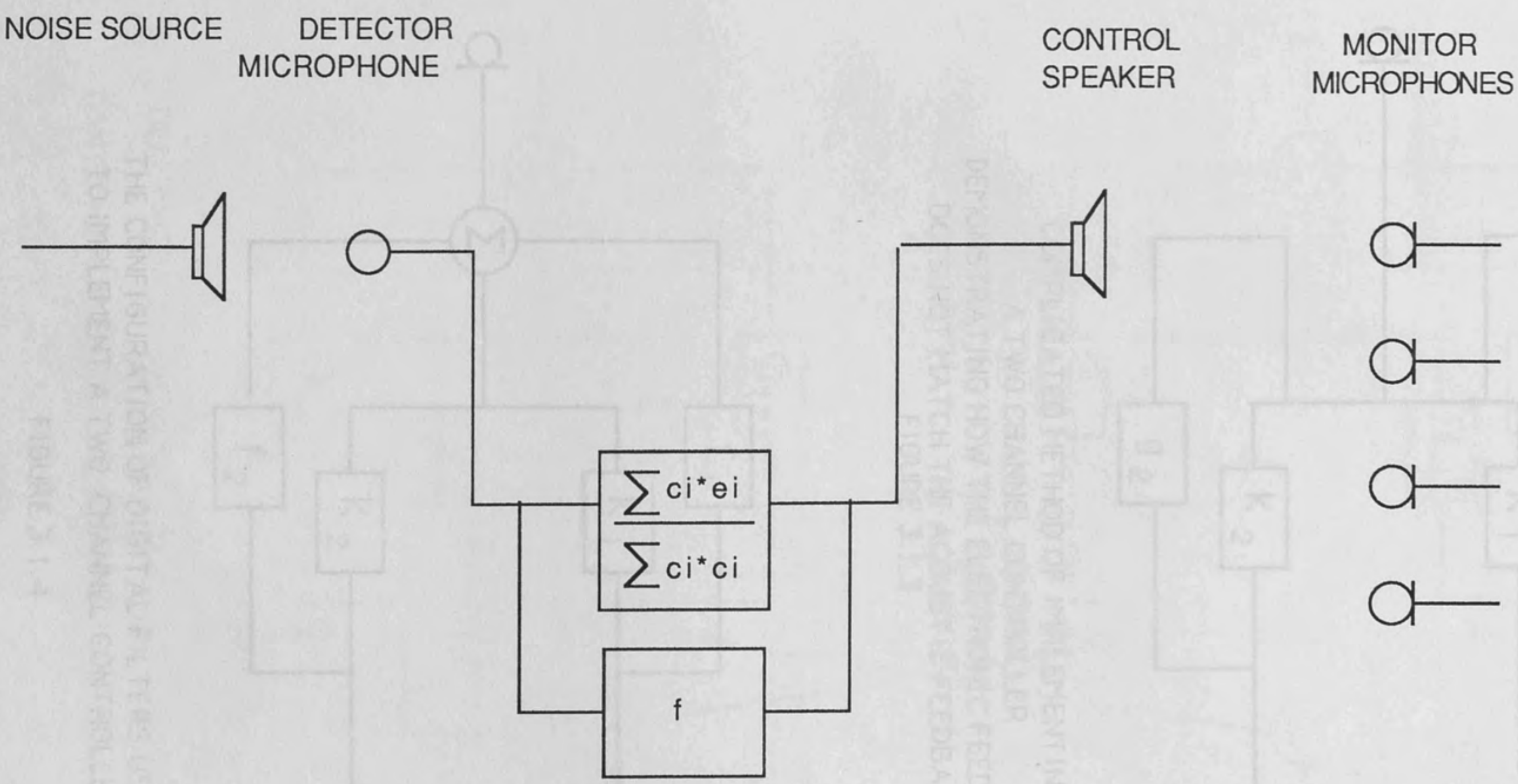
FIGURE 2.3.7



THE GENERAL FORMAT FOR AN ACTIVE NOISE CONTROL SYSTEM

SHOWING THE RELEVANT TRANSFER FUNCTIONS

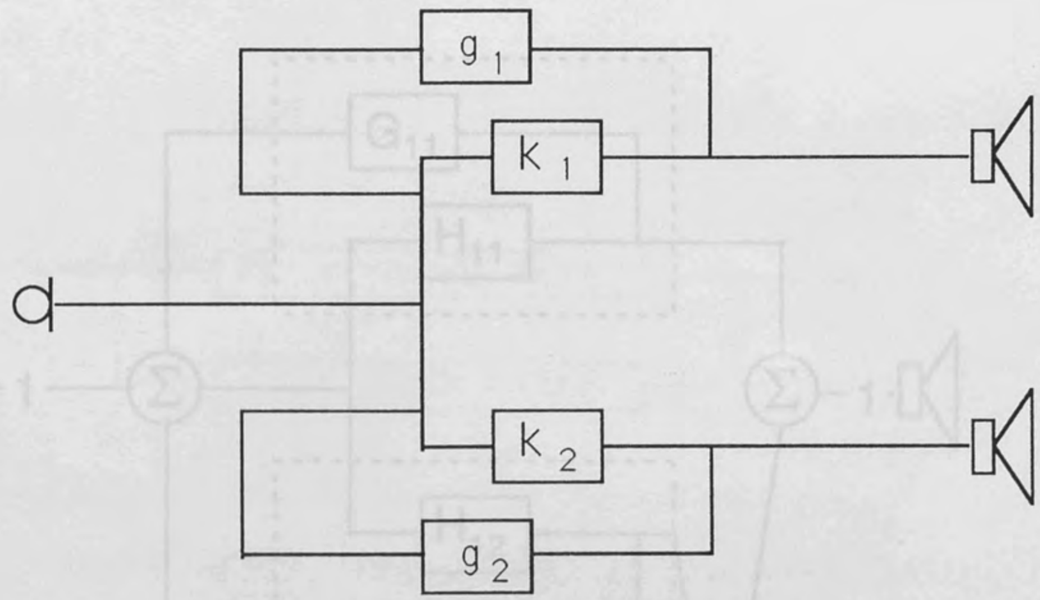
FIGURE 3.1.1



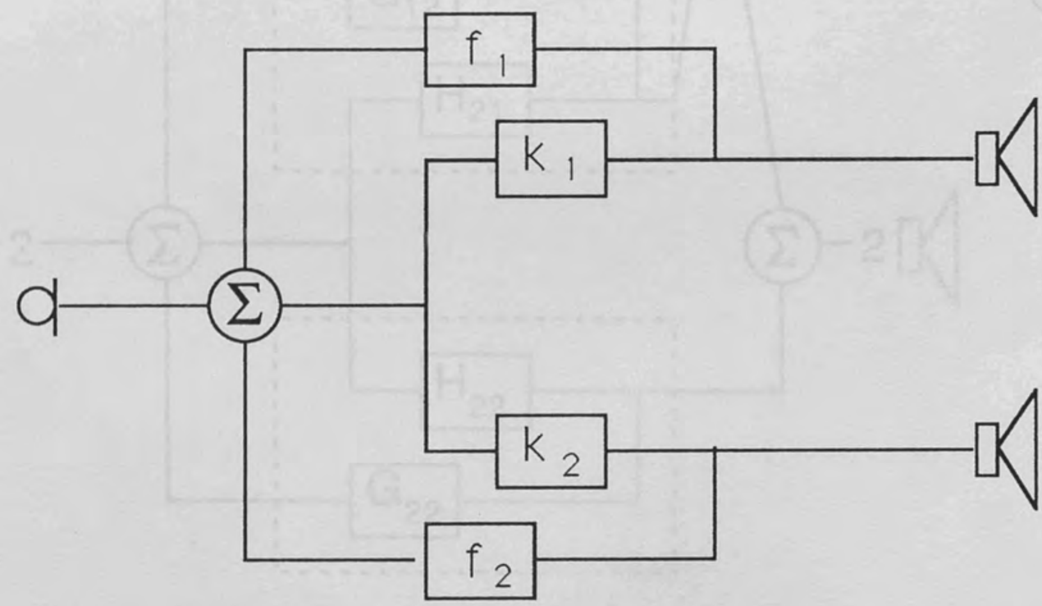
ACTIVE CONTROL SYSTEM TO CONTROL THE FIELD AT A NUMBER OF LOCATIONS

THE DESIRED RESPONSE OF THE DIGITAL CONTROLLER IS SHOWN

FIGURE 3.1.2

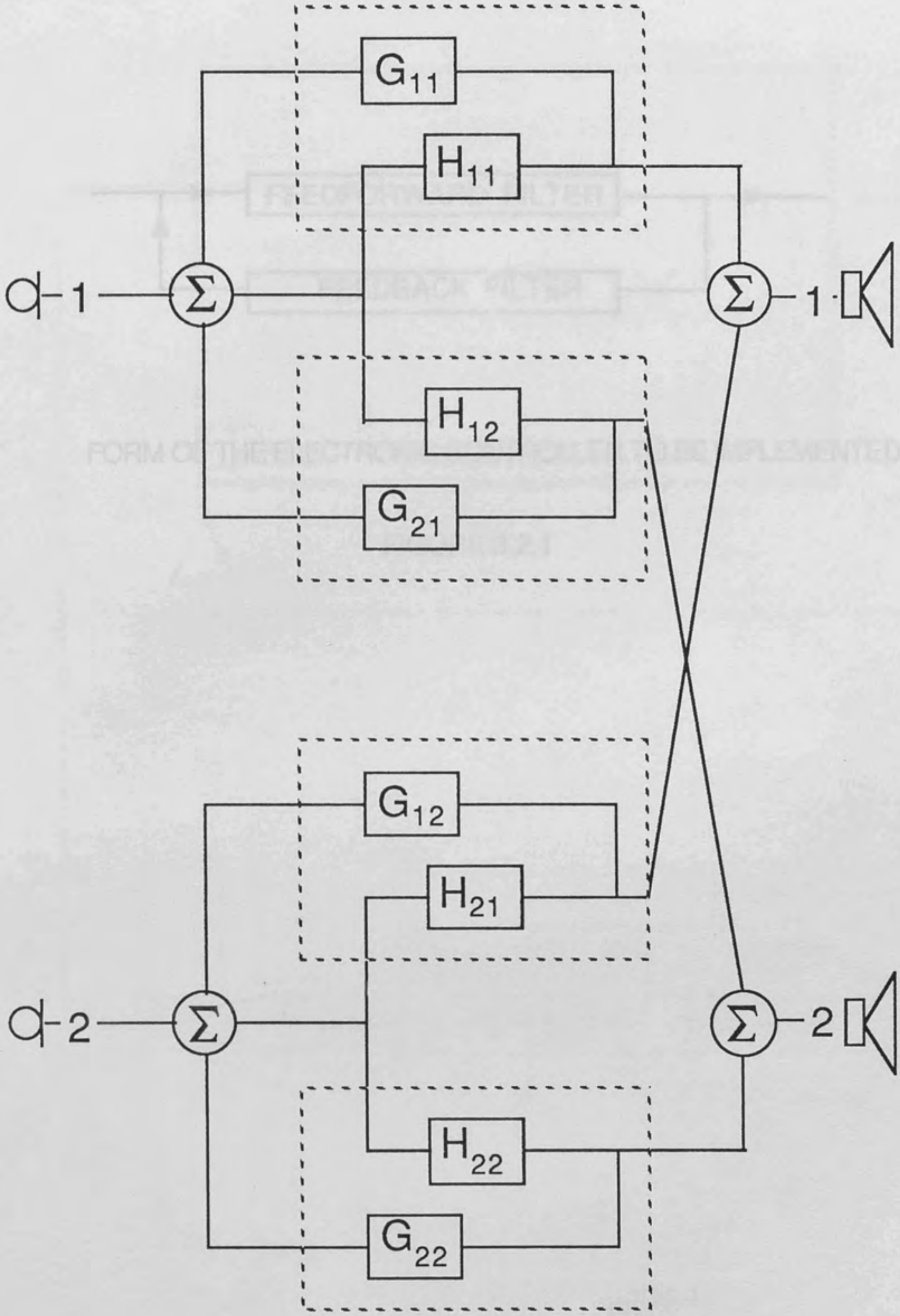


COMPLICATED METHOD OF IMPLEMENTING
A TWO CHANNEL CONTROLLER
DEMONSTRATING HOW THE ELECTRONIC FEEDBACK
DOES NOT MATCH THE ACOUSTIC FEEDBACK
FIGURE 3.1.3



THE CONFIGURATION OF DIGITAL FILTERS USED
TO IMPLEMENT A TWO CHANNEL CONTROLLER

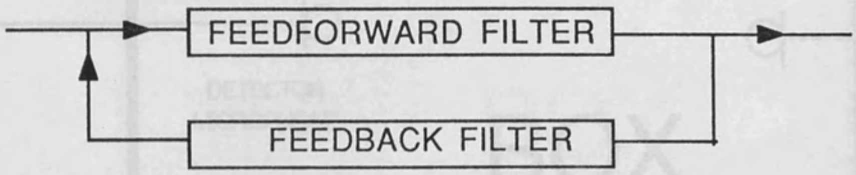
FIGURE 3.1.4



DEMONSTRATION OF HOW A SINGLE CHANNEL CONTROLLER

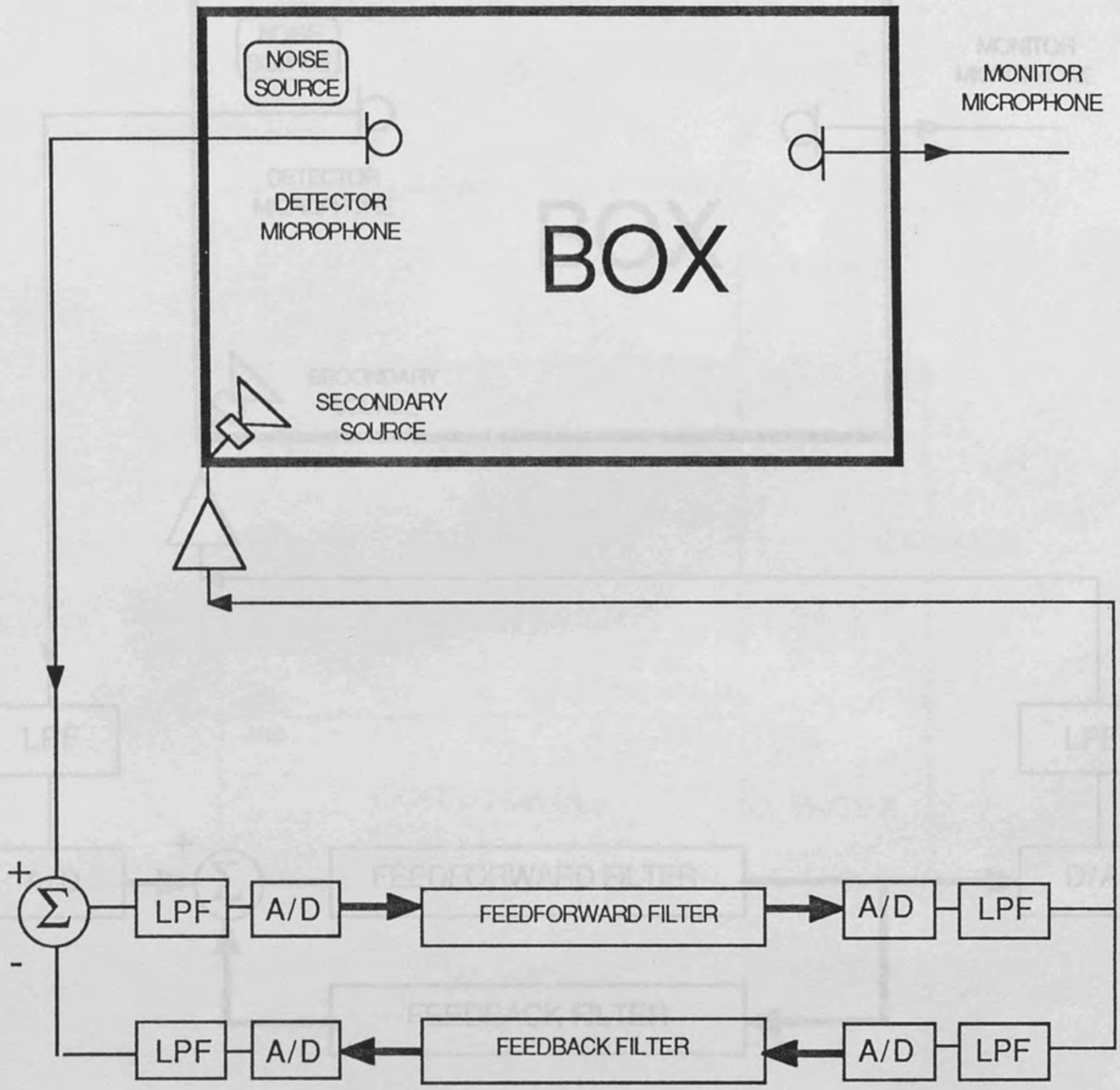
CAN BE REPEATED TO IMPLEMENT MULTI-CHANNEL CONTROL

FIGURE 3.1.5



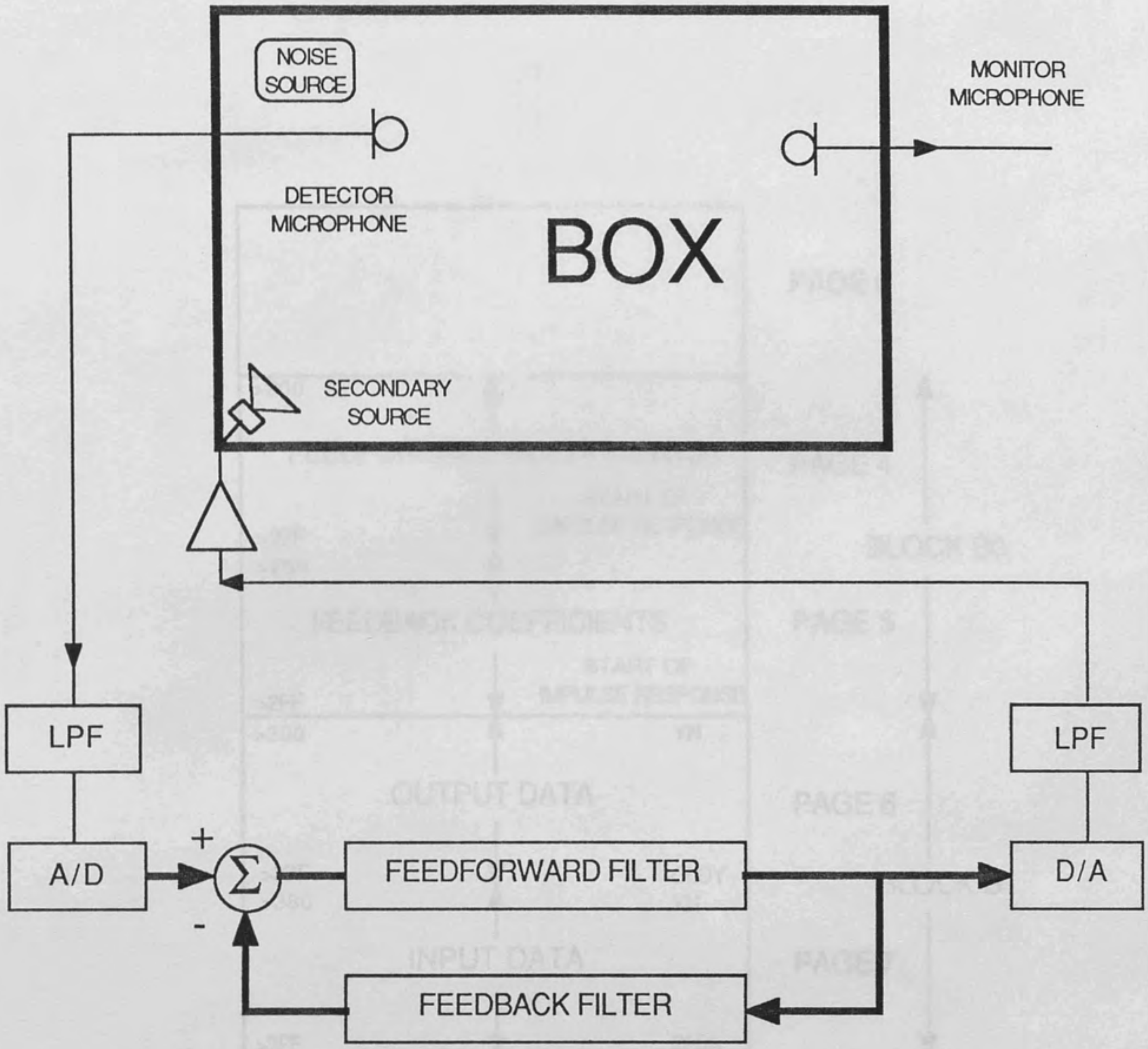
FORM OF THE ELECTRONIC CONTROLLER TO BE IMPLEMENTED

FIGURE 3.2.1

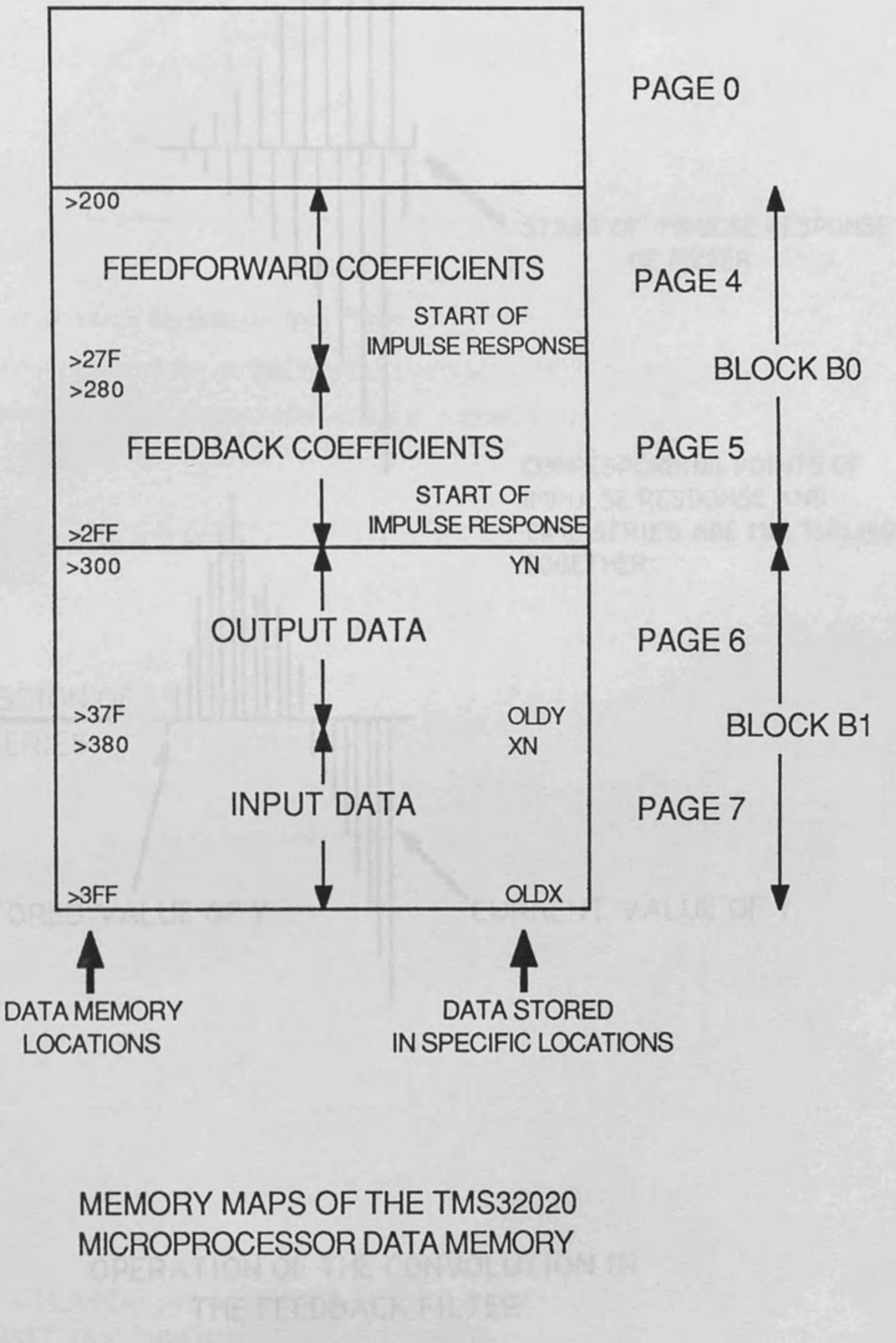


POSSIBLE REALISATION OF AN
ACTIVE NOISE CONTROL SYSTEM
OPERATING IN AN ENCLOSURE

FIGURE 3.2.2

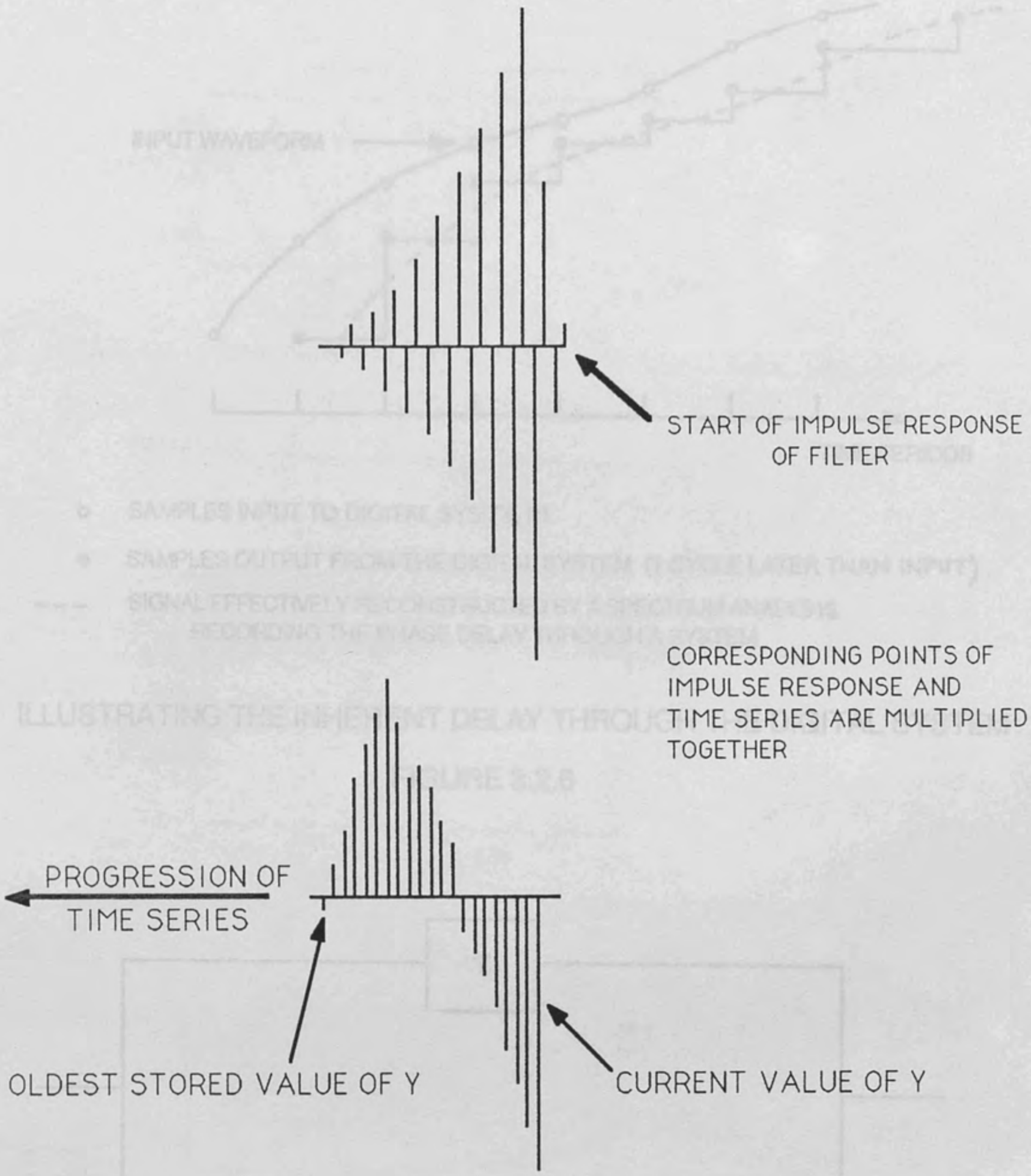


ACTIVE NOISE CONTROL SYSTEM
OPERATING IN AN ENCLOSURE
FIGURE 3.2.3



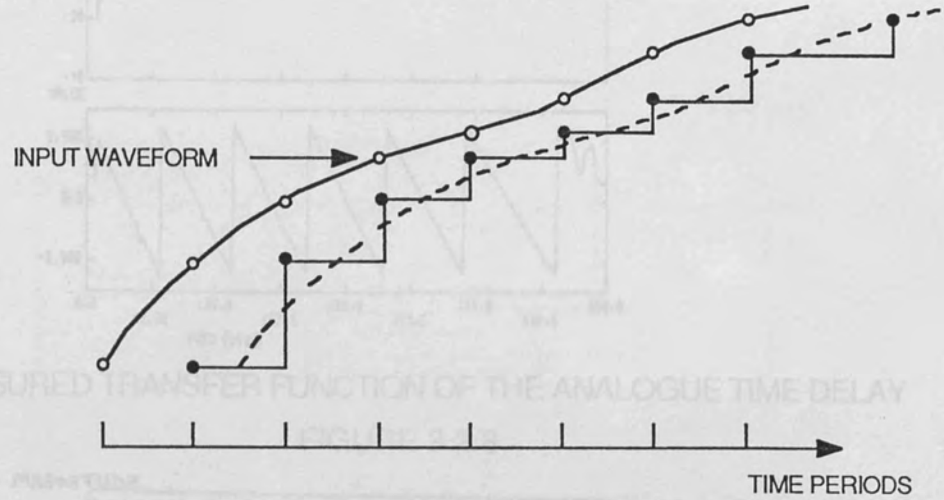
MEMORY MAPS OF THE TMS32020
MICROPROCESSOR DATA MEMORY

FIGURE 3.2.4



OPERATION OF THE CONVOLUTION IN THE FEEDBACK FILTER

FIGURE 3.2.5



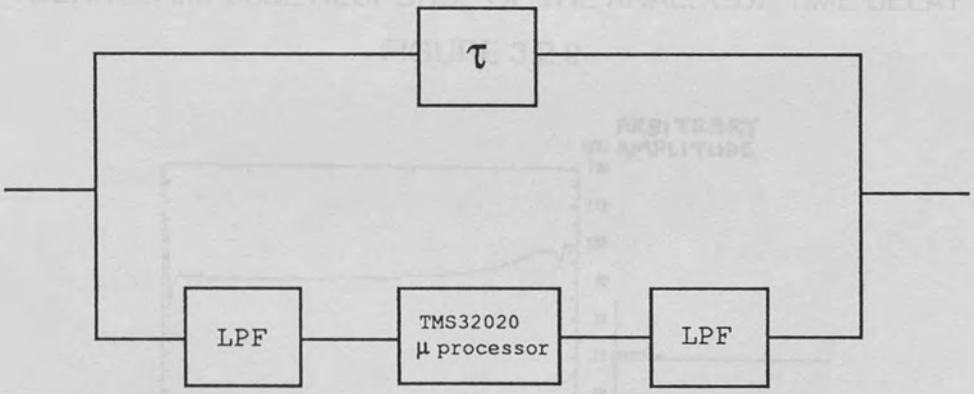
o SAMPLES INPUT TO DIGITAL SYSTEM

• SAMPLES OUTPUT FROM THE DIGITAL SYSTEM (1 CYCLE LATER THAN INPUT)

--- SIGNAL EFFECTIVELY RECONSTRUCTED BY A SPECTRUM ANALYSIS RECORDING THE PHASE DELAY THROUGH A SYSTEM

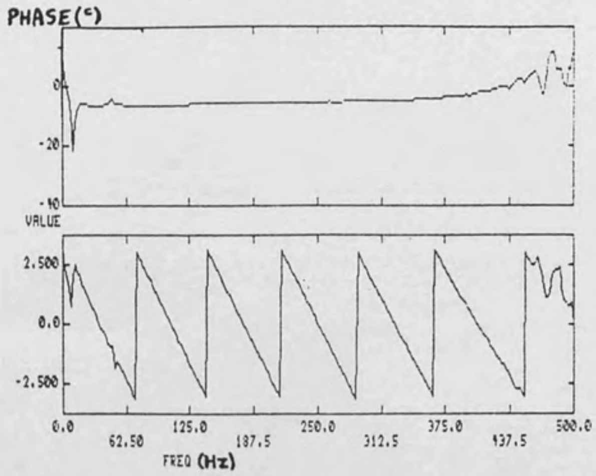
ILLUSTRATING THE INHERENT DELAY THROUGH THE DIGITAL SYSTEM

FIGURE 3.2.6



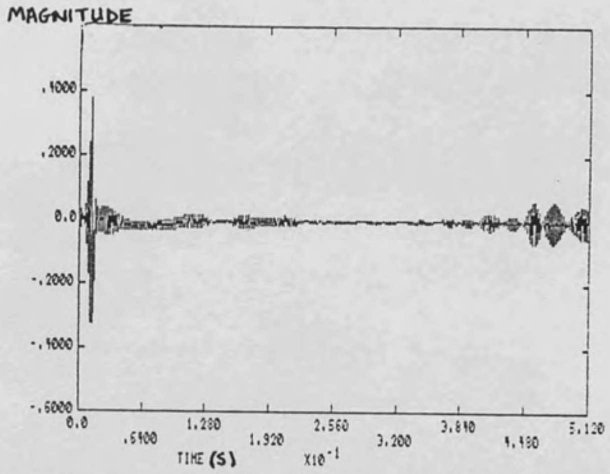
SIMULATION OF AN ANALOGUE TIME DELAY DEVICE TO TEST THE OPERATION OF A DIGITAL FILTER

FIGURE 3.2.7



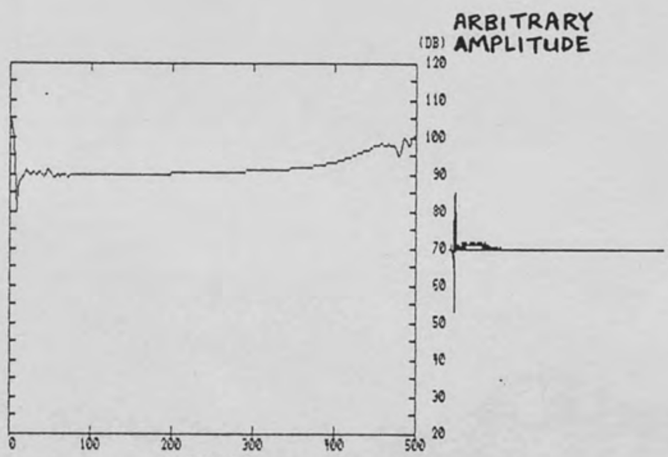
MEASURED TRANSFER FUNCTION OF THE ANALOGUE TIME DELAY

FIGURE 3.2.8



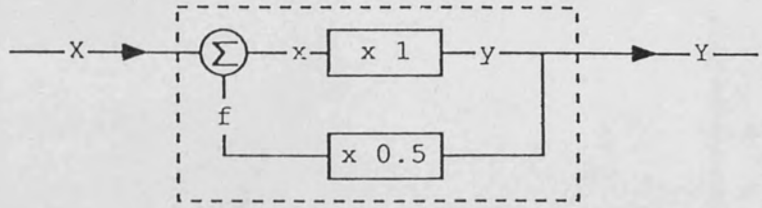
DERIVED IMPULSE RESPONSE OF THE ANALOGUE TIME DELAY

FIGURE 3.2.9



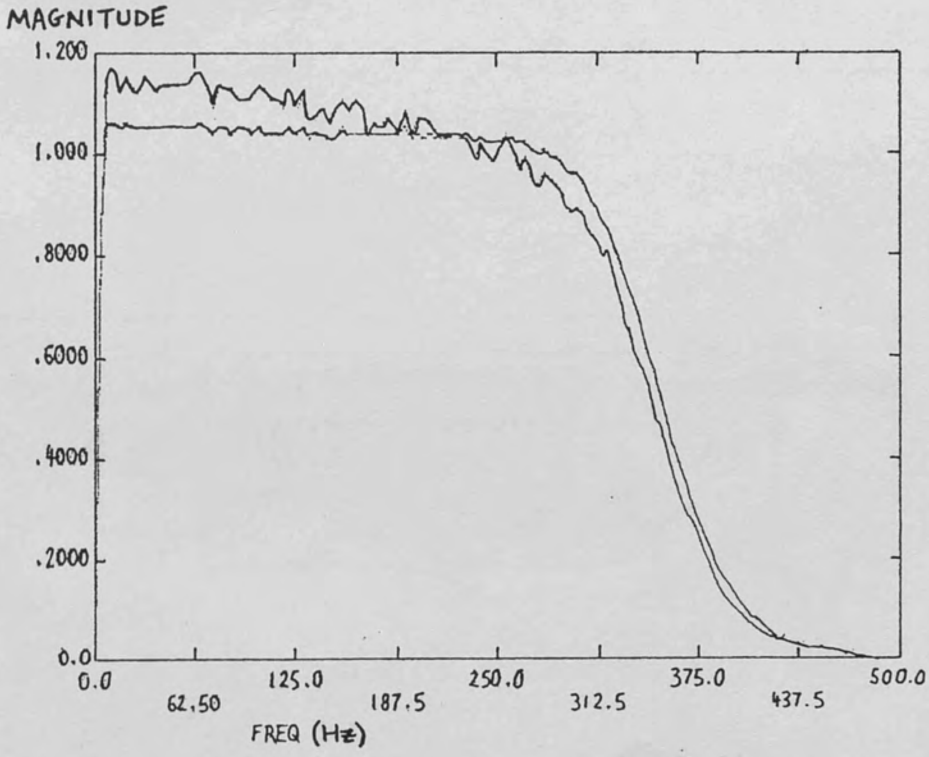
128 POINT FIR FILTER MODELLING
THE ANALOGUE TIME DELAY AND LOW PASS FILTERS

FIGURE 3.2.10



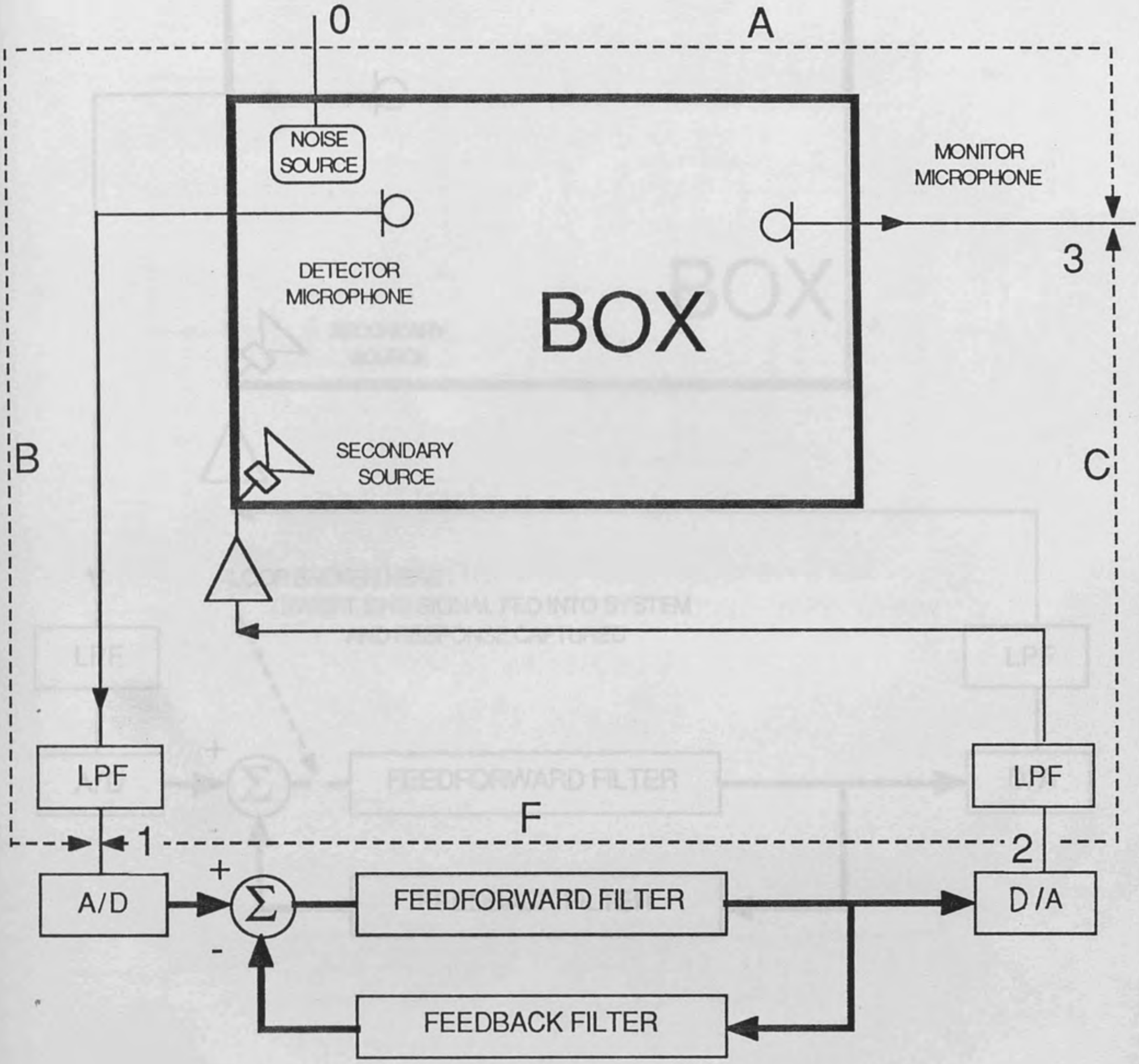
TESTING THE CLOSED LOOP OF THE CONTROLLER
USING FILTERS OF SIMPLE-KNOWN RESPONSES

FIGURE 3.2.11



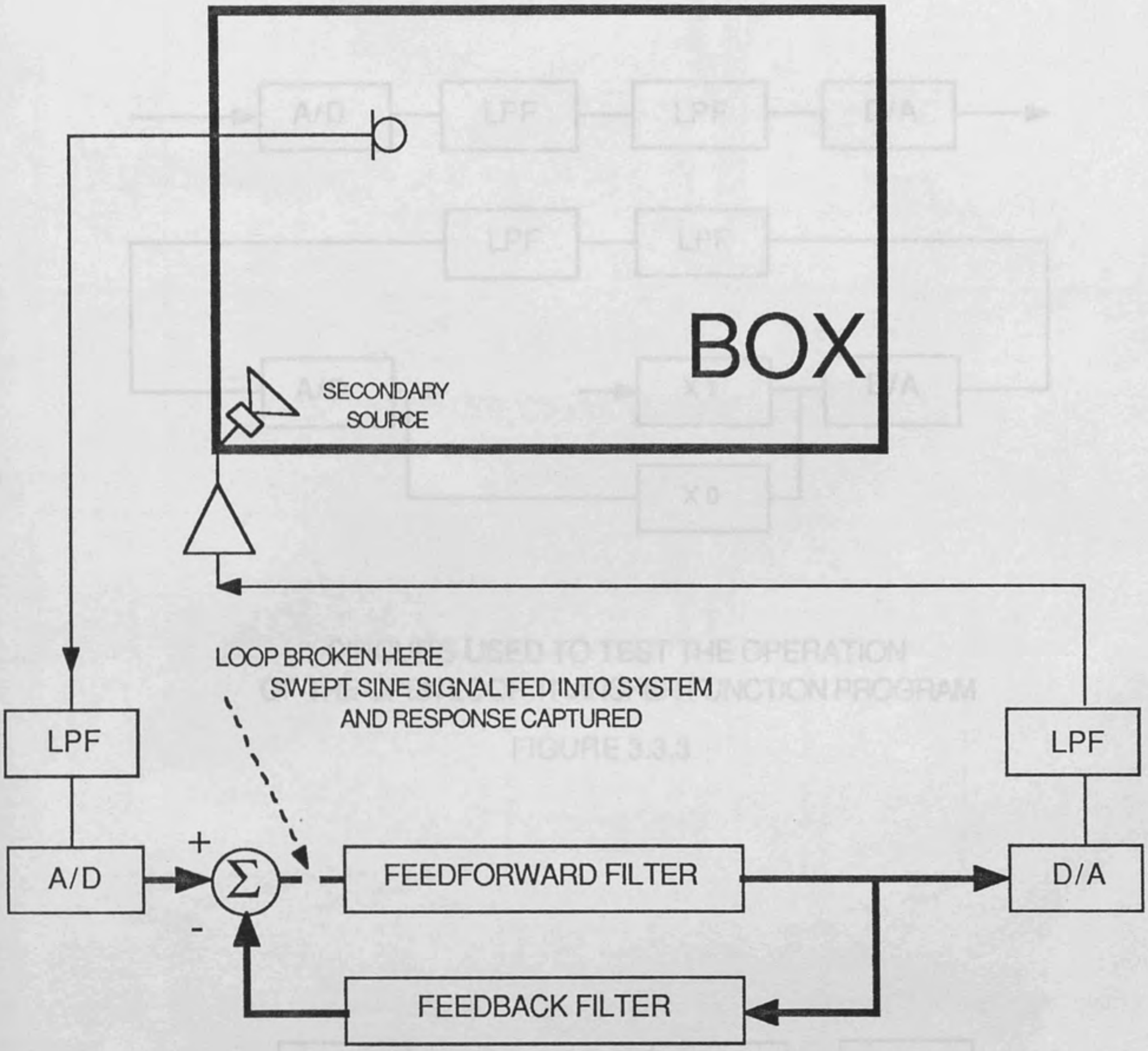
TRANSFER FUNCTIONS OF THE FILTER OF
FIGURE 3.2.11 AND THAT OF THE ANALOGUE FILTERS ALONE

FIGURE 3.2.12



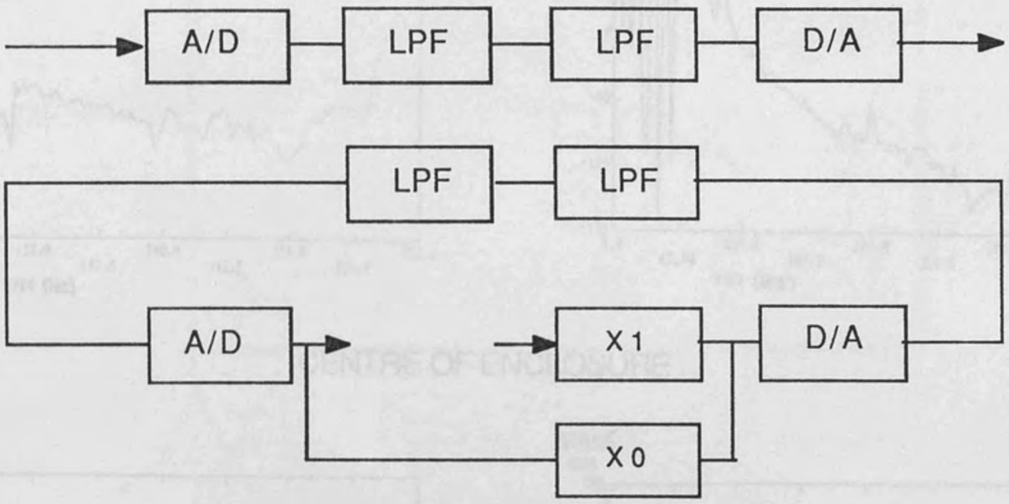
ACTIVE NOISE CONTROL SYSTEM
OPERATING IN AN ENCLOSURE
SHOWING RELEVANT TRANSFER FUNCTIONS
AND POSITIONAL NOTATION NUMBERS

FIGURE 3.3.1



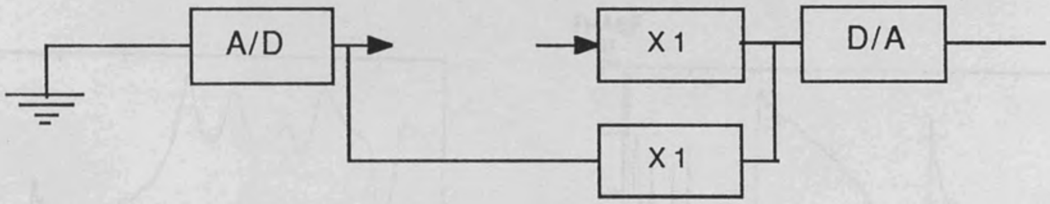
SYSTEM TO MEASURE THE
OPEN LOOP TRANSFER FUNCTION
ROUND AN ACTIVE CONTROL SYSTEM

FIGURE 3.3.2



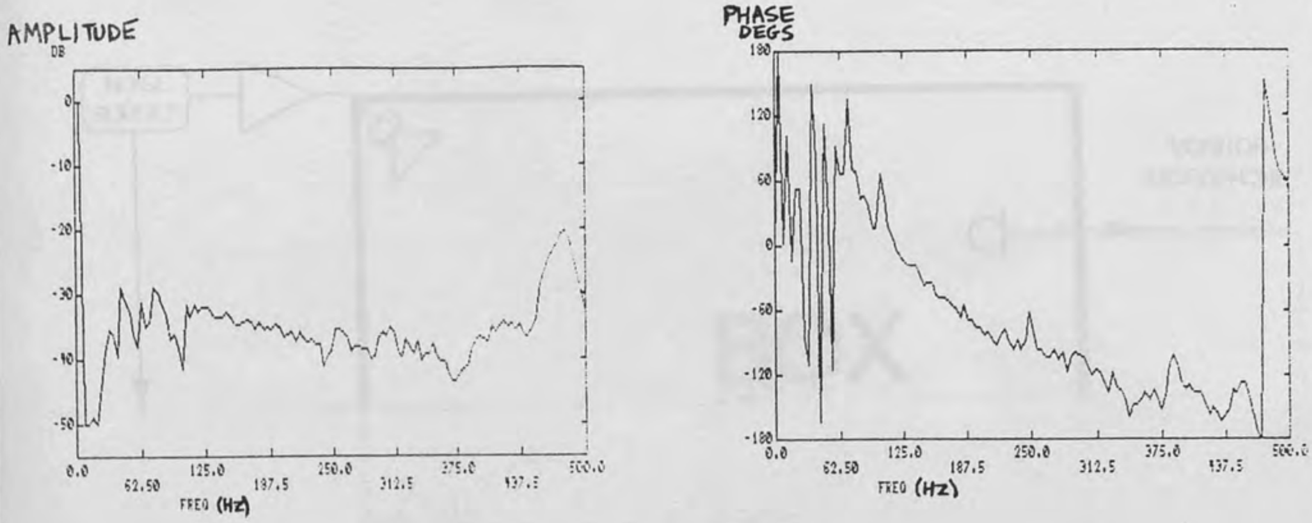
CIRCUITS USED TO TEST THE OPERATION
OF THE OPEN LOOP TRANSFER FUNCTION PROGRAM

FIGURE 3.3.3

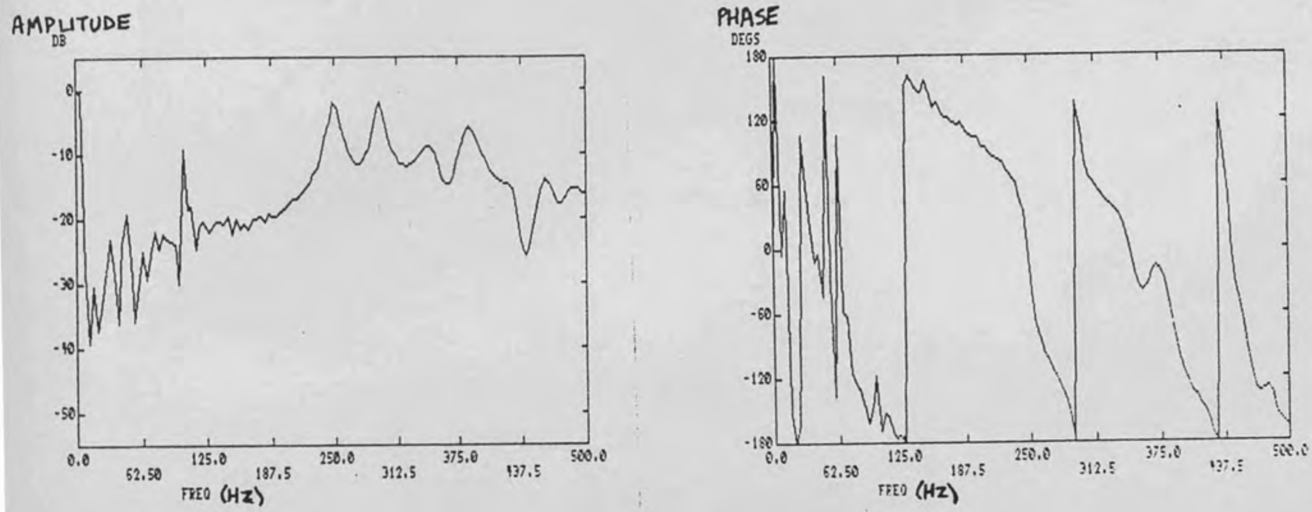


TEST TO VERIFY THE INTERNAL OPERATION
OF THE OPEN LOOP TRANSFER PROGRAM

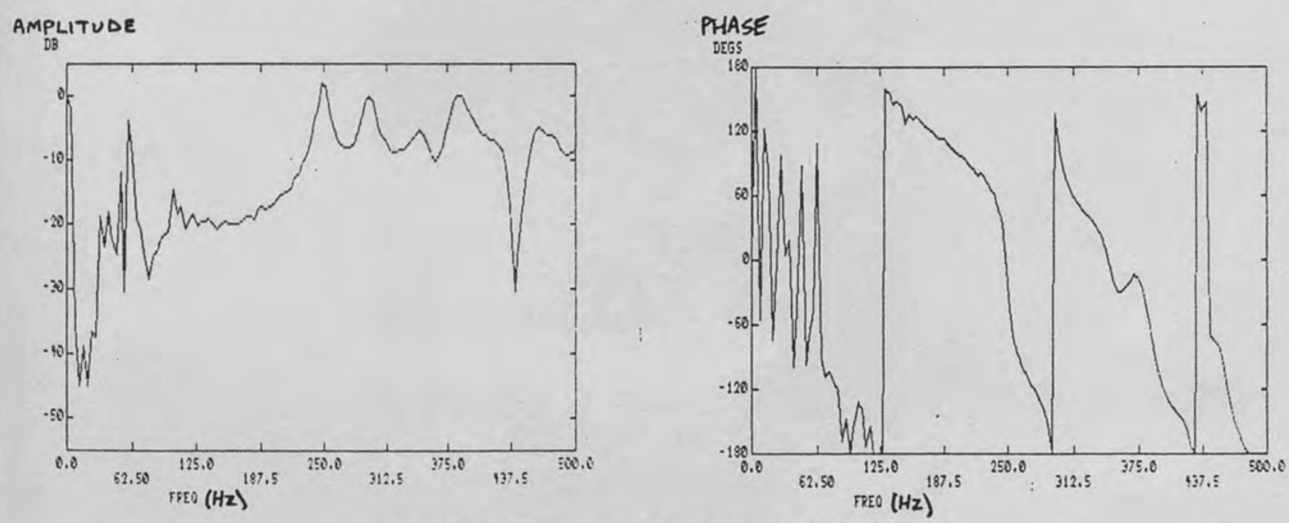
FIGURE 3.3.4



CENTRE OF ENCLOSURE

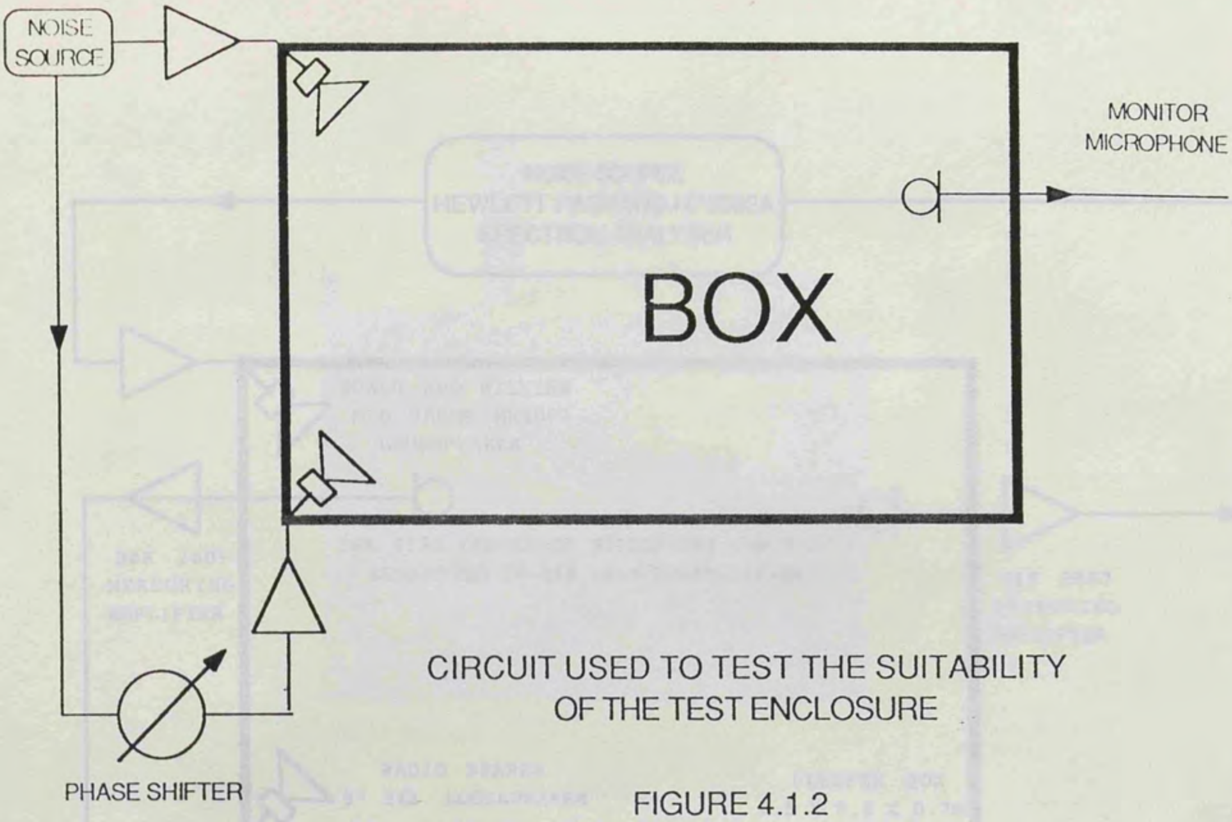


MID WAY ALONG A DIAGONAL IN THE OPPOSITE CORNER TO THE SPEAKER



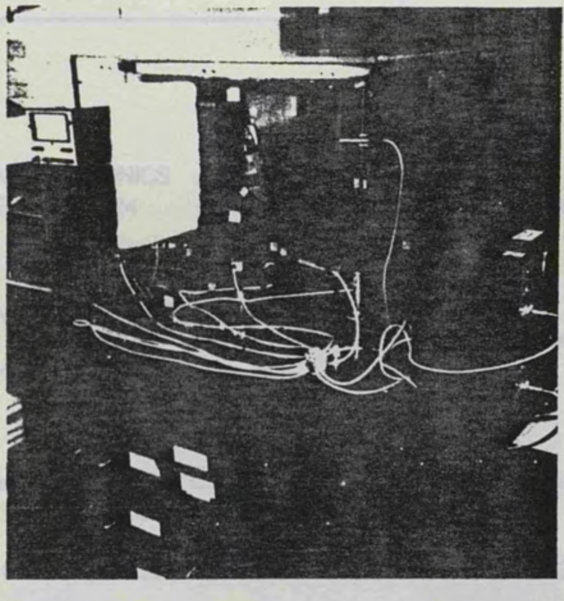
AT OPPOSITE CORNER TO THE SPEAKER

FIGURE 4.1.1 REPRESENTATIVE PLOTS OF THE SOUND FIELD



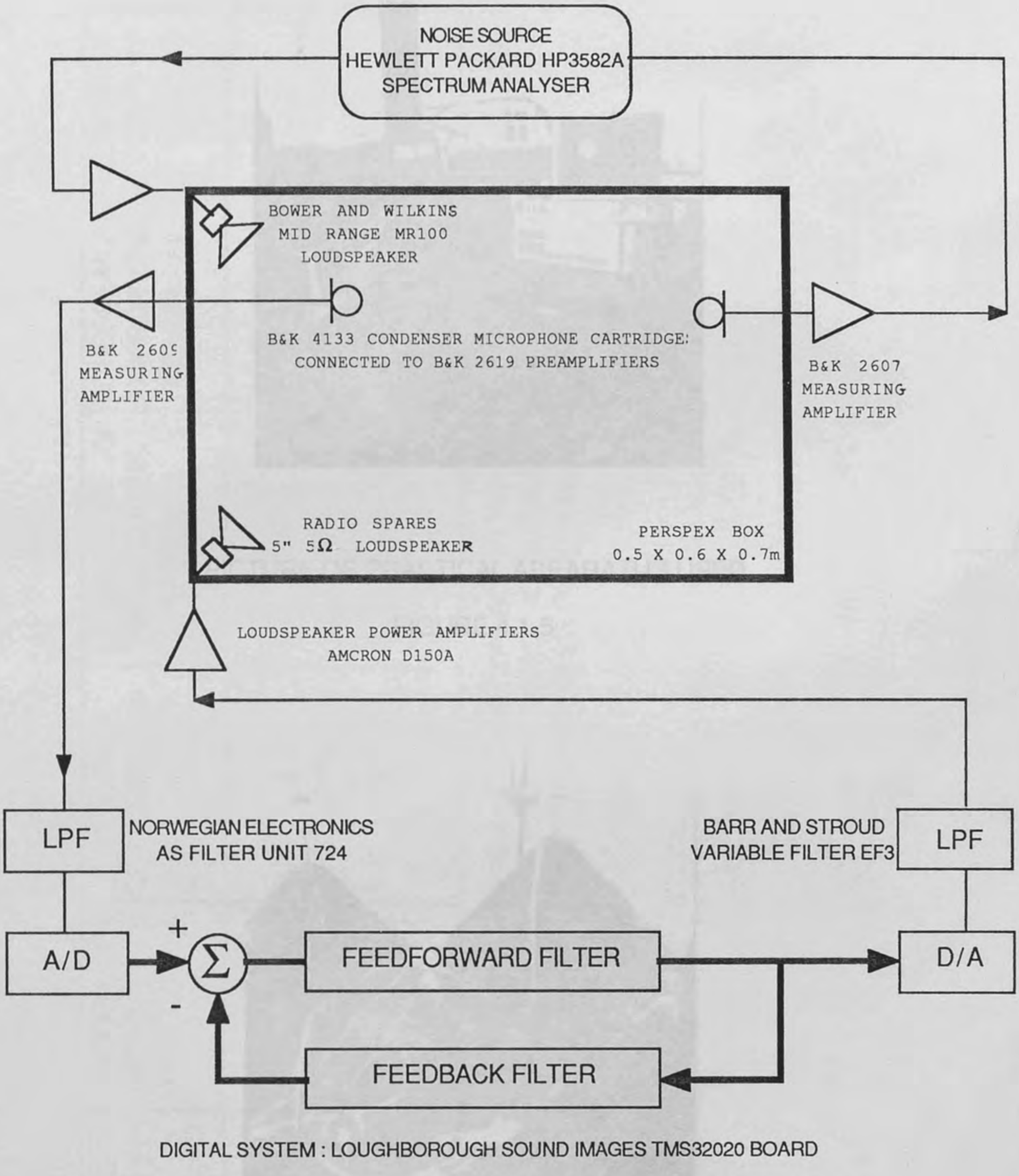
CIRCUIT USED TO TEST THE SUITABILITY
OF THE TEST ENCLOSURE

FIGURE 4.1.2

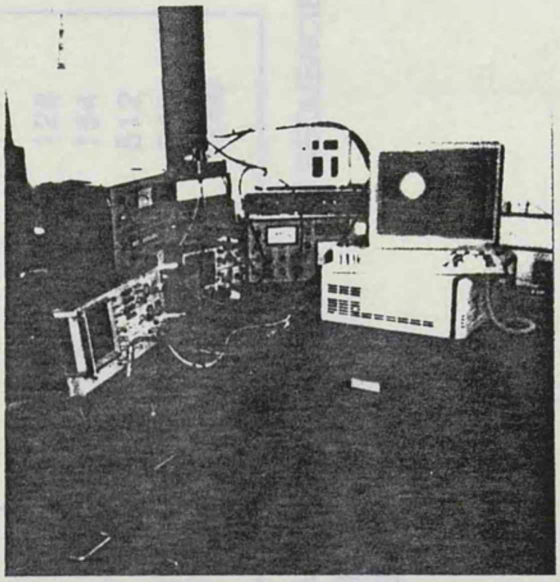


PICTURE OF TEST ENCLOSURE
SHOWING PASSIVE DAMPING
ON TWO FACES

FIGURE 4.1.3

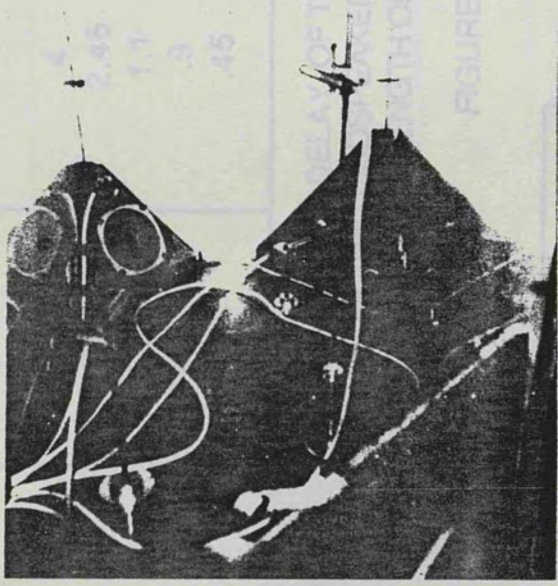


ACTIVE NOISE CONTROL SYSTEM
OPERATING IN AN ENCLOSURE
SHOWING PRACTICAL APPARATUS USED
FIGURE 4.1.4



PICTURE OF PRACTICAL APPARATUS USED

FIGURE 4.1.5



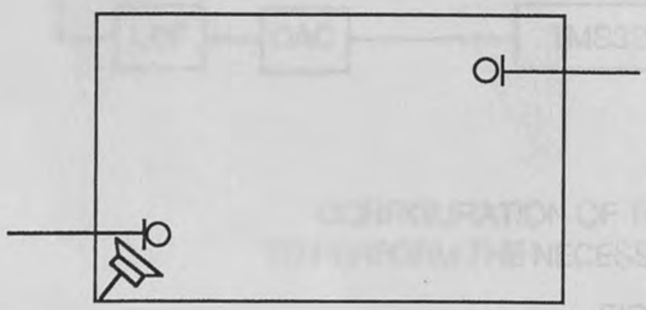
PICTURE OF TRANSDUCERS INSIDE THE TEST ENCLOSURE

FIGURE 4.1.7

SAMPLING FREQUENCY (Hz)	CUT OFF FREQ OF LOW PASS FILTERS (Hz)	GROUP DELAY OF FILTERS (mS)	ESTIMATED DELAY OF ELECTRONIC PATH (mS)	LOUDSPEAKER-MICROPHONE SPACING (m)	NUMBER OF POINTS IN FIR FILTER
1000	350	4	6	2	128
1200	450	2.45	4.3	1.5	154
4000	1500	1.1	2.4	.8	512
5000	1800	.9	2.1	.7	640
10000	3500	.45	1.6	.5	1280

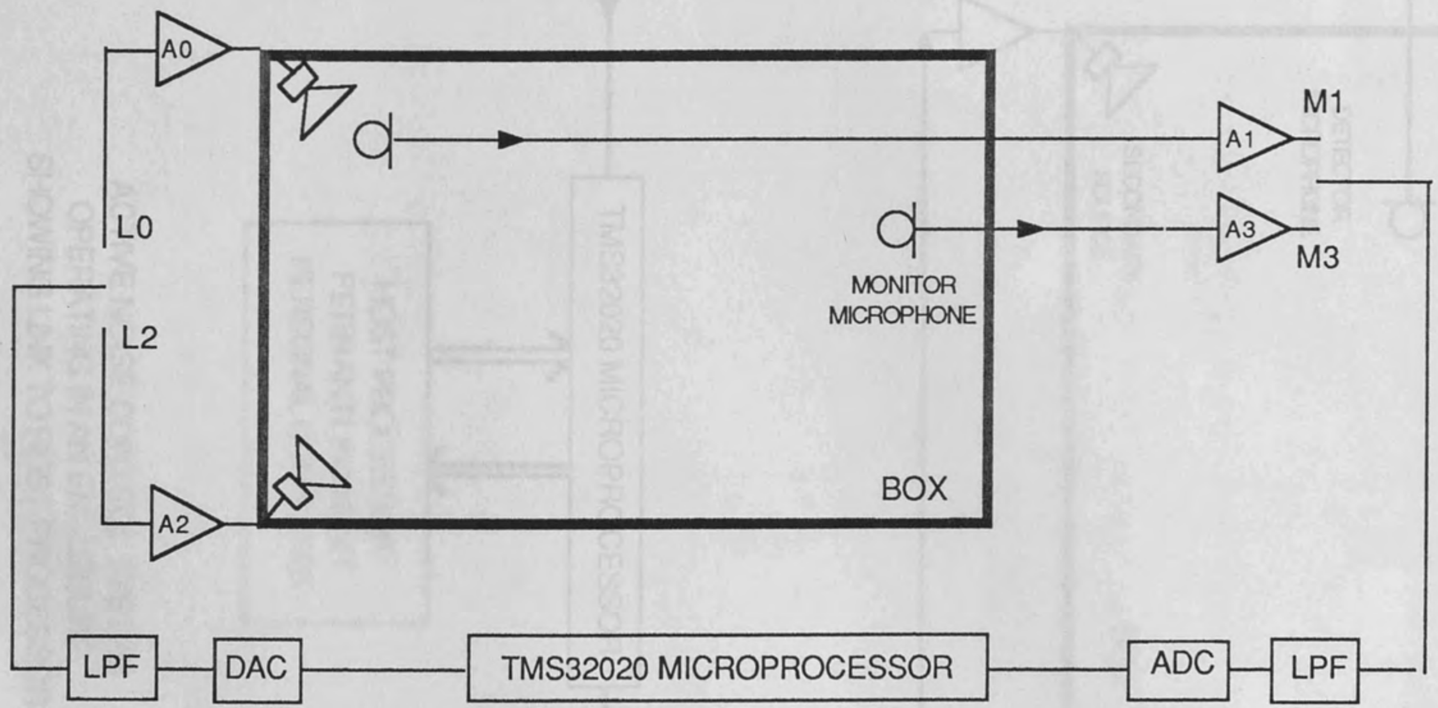
ILLUSTRATIVE DATA OF THE ELECTRONIC DELAY OF THE CONTROL SYSTEM AT VARIOUS SAMPLING FREQUENCIES SHOWING THE LOUDSPEAKER-MICROPHONE SPACING REQUIRED AND THE LENGTH OF THE FIR FILTERS NEEDED

FIGURE 4.1.6



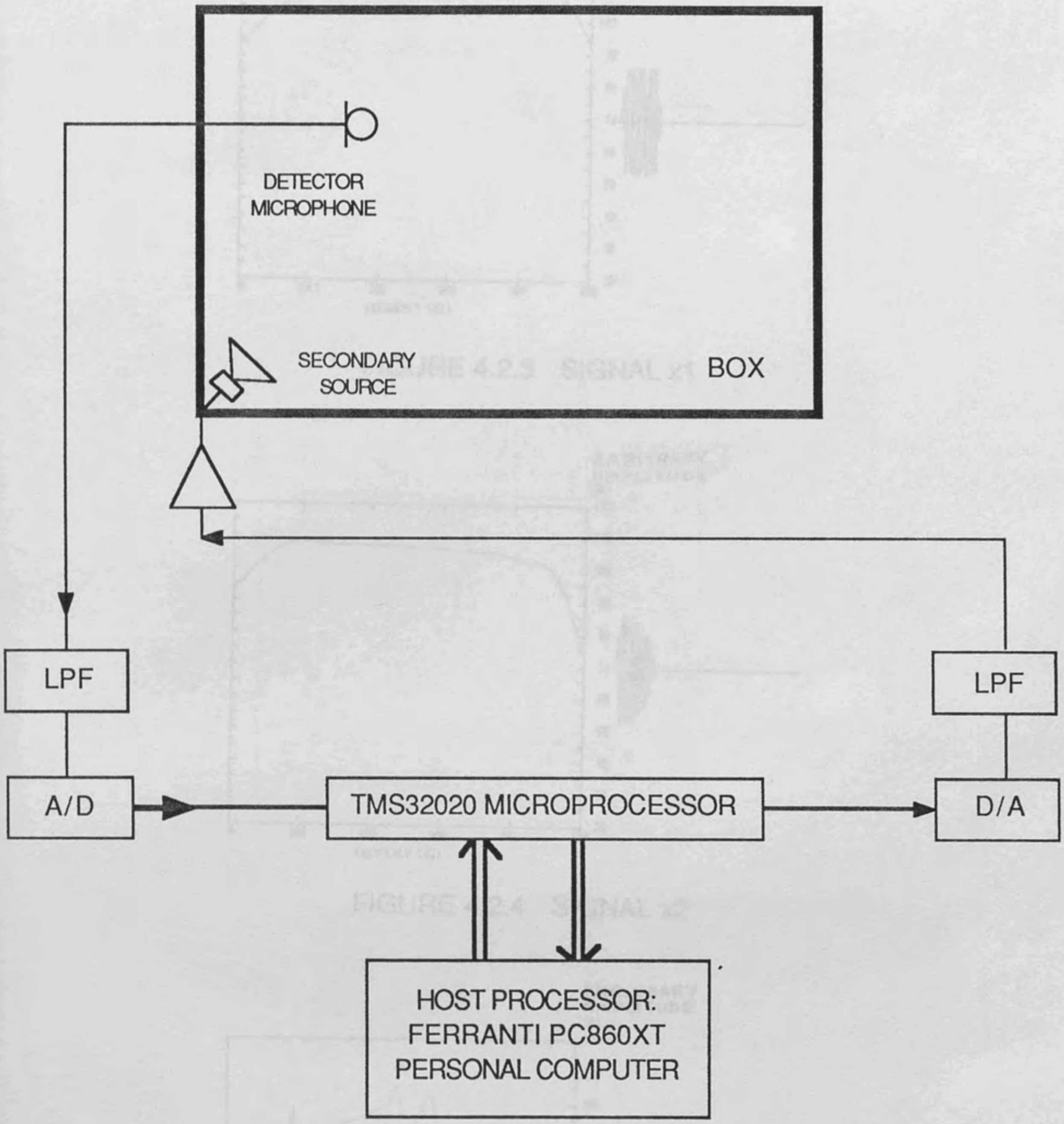
SIDE PROJECTION OF THE TEST ENCLOSURE

FIGURE 4.1.8



CONFIGURATION OF THE CONTROL SYSTEM APPARATUS
TO PERFORM THE NECESSARY MEASUREMENTS ON THE SYSTEM

FIGURE 4.2.1



ACTIVE NOISE CONTROL SYSTEM
OPERATING IN AN ENCLOSURE
SHOWING LINK TO HOST PROCESSOR

FIGURE 4.2.2

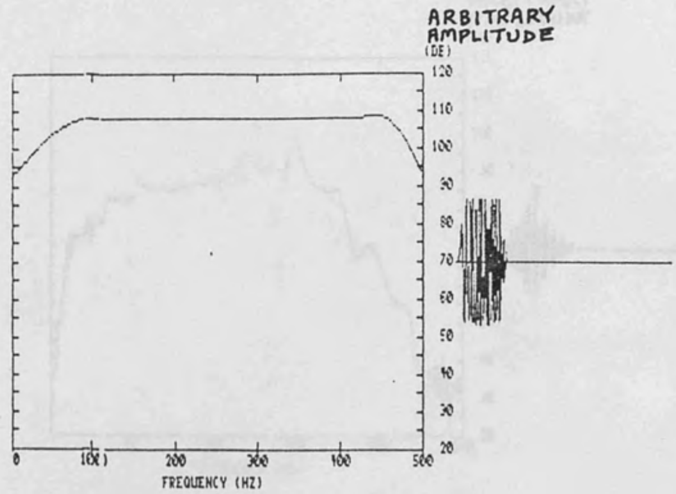


FIGURE 4.2.3 SIGNAL x1

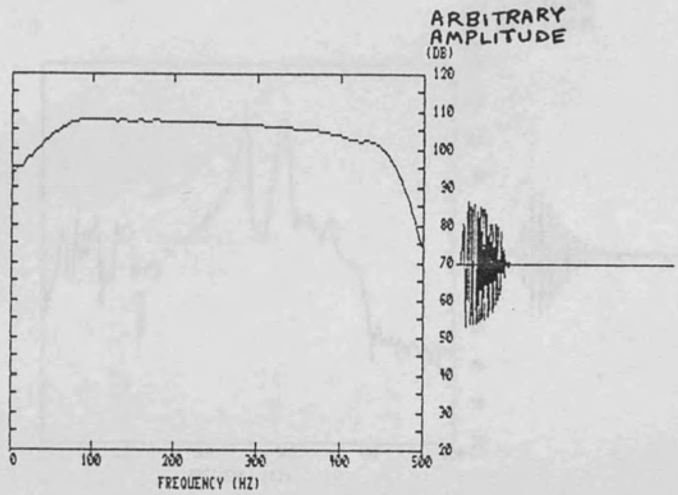


FIGURE 4.2.4 SIGNAL x2

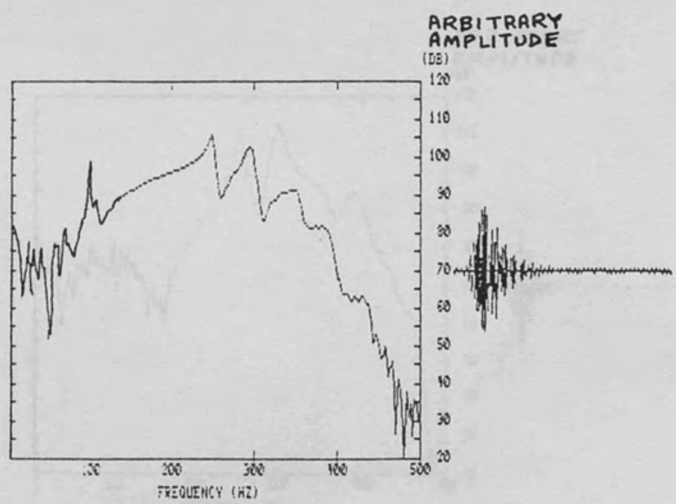


FIGURE 4.2.5 SIGNAL y10

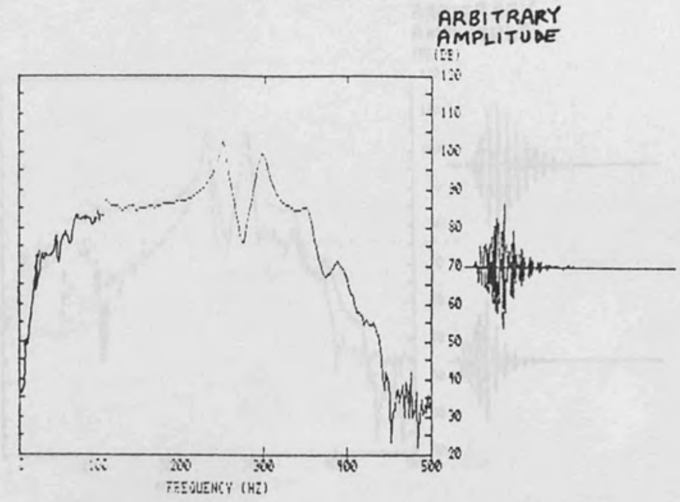


FIGURE 4.2.6 SIGNAL y30

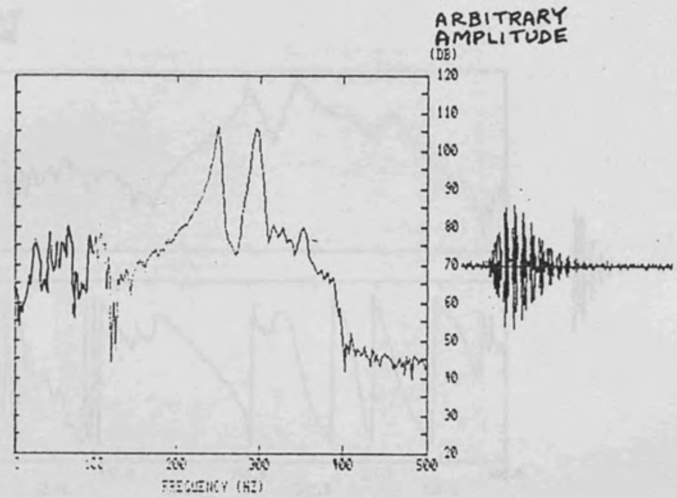


FIGURE 4.2.7 SIGNAL y32

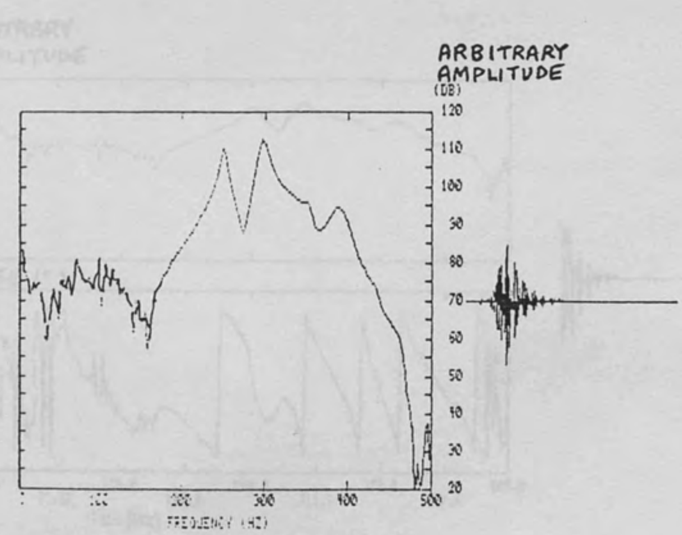


FIGURE 4.2.8 SIGNAL y12

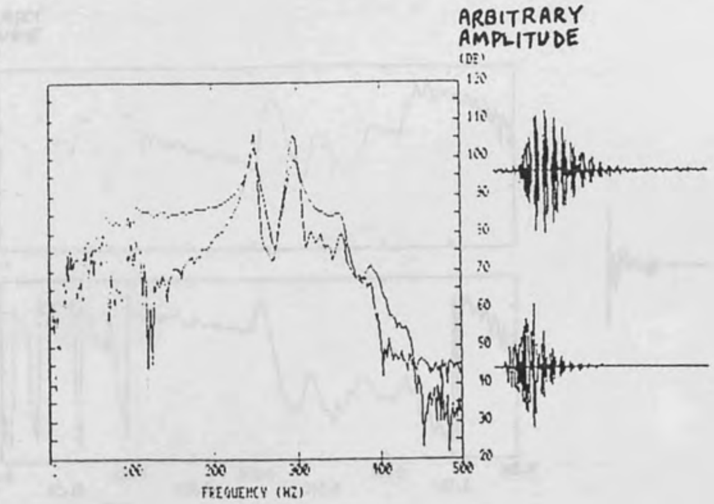


FIGURE 4.2.9 SIGNALS y30 and y32

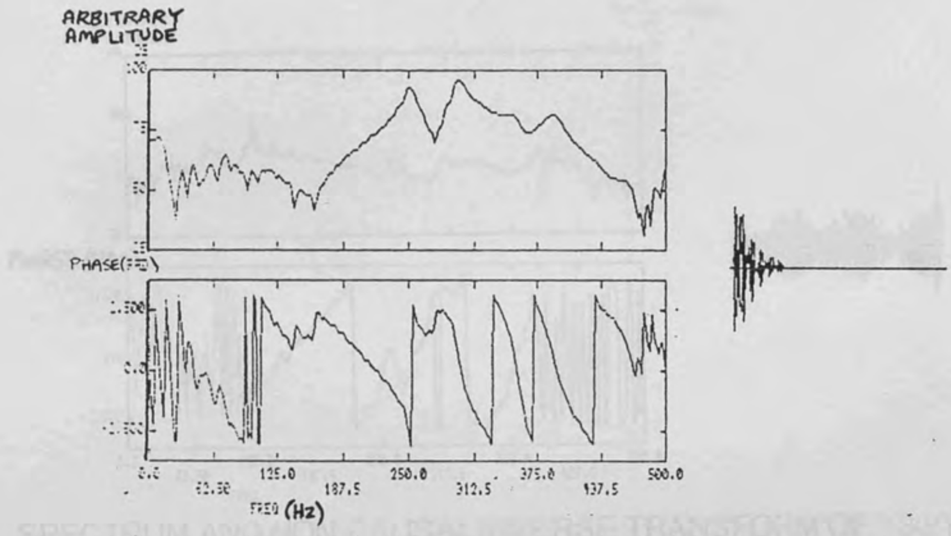


FIGURE 4.2.10 DERIVED FEEDBACK FILTER 128 POINT FIR

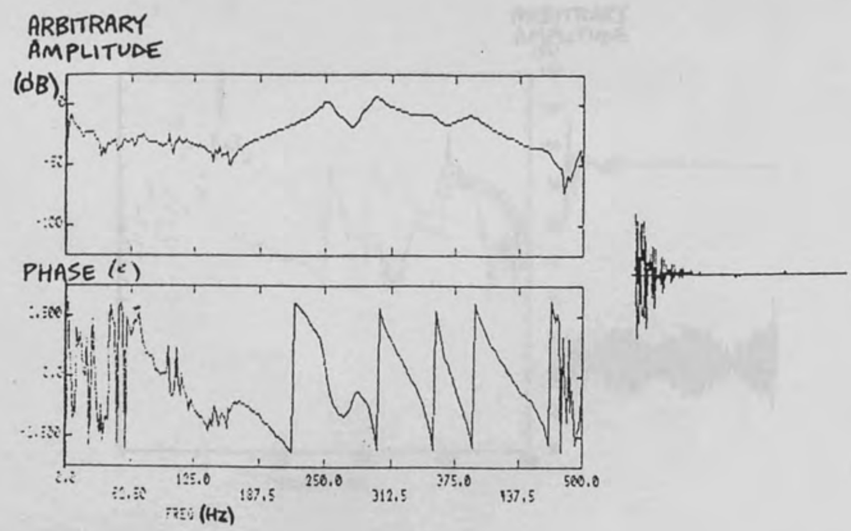
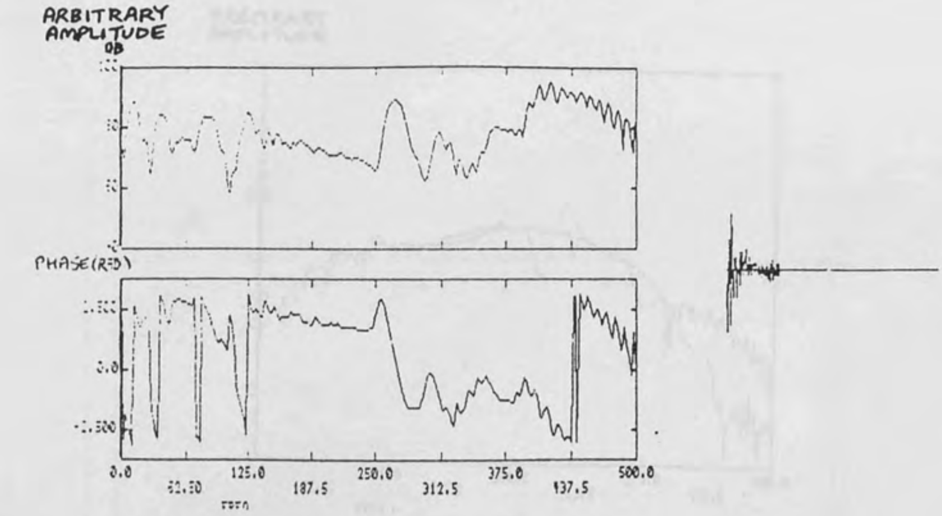
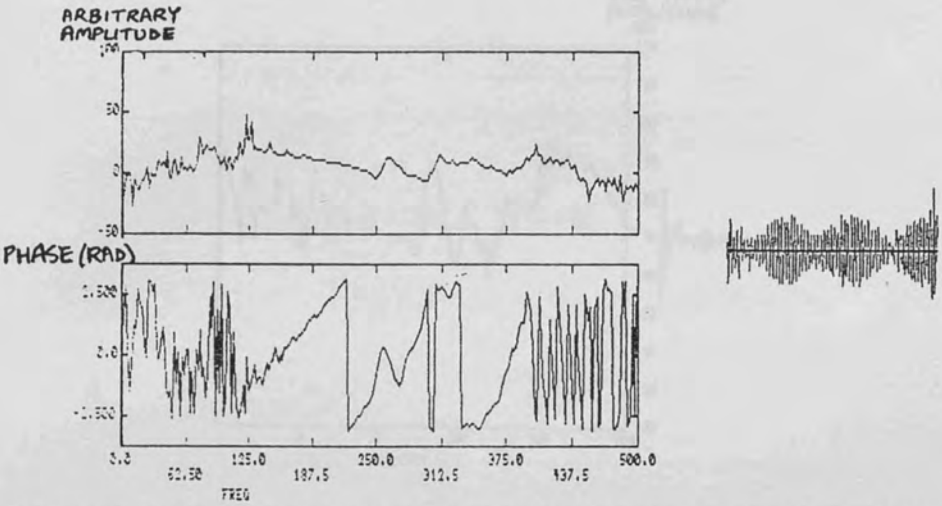


FIGURE 4.2.11 SPECTRUM AND INVERSE TRANSFORM OF Y12/X2



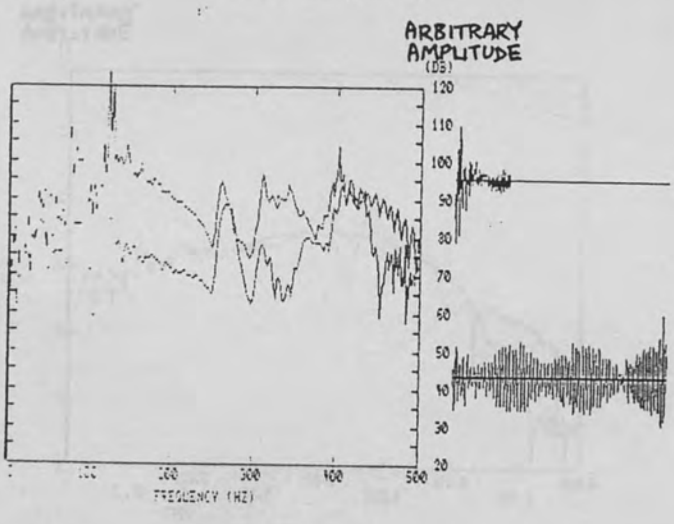
DERIVED FEEDFORWARD FILTER 128 POINT FIR

FIGURE 4.2.12



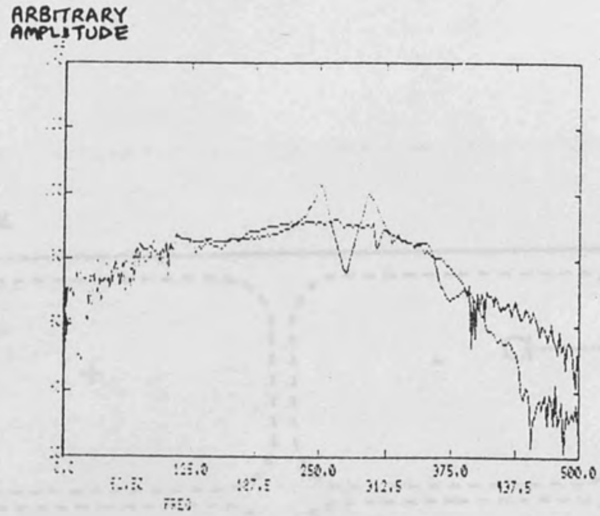
SPECTRUM AND NON-CAUSAL INVERSE TRANSFORM OF Y30/Y32

FIGURE 4.2.13



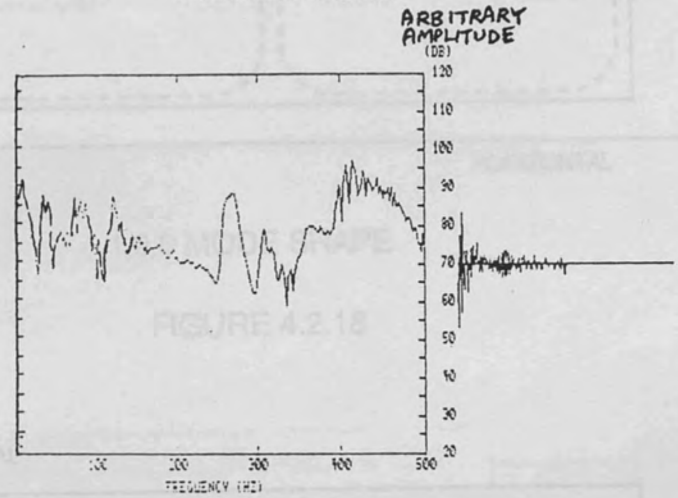
AMPLITUDE SPECTRUM OF FEEDFORWARD FILTER AND Y30/Y32

FIGURE 4.2.14



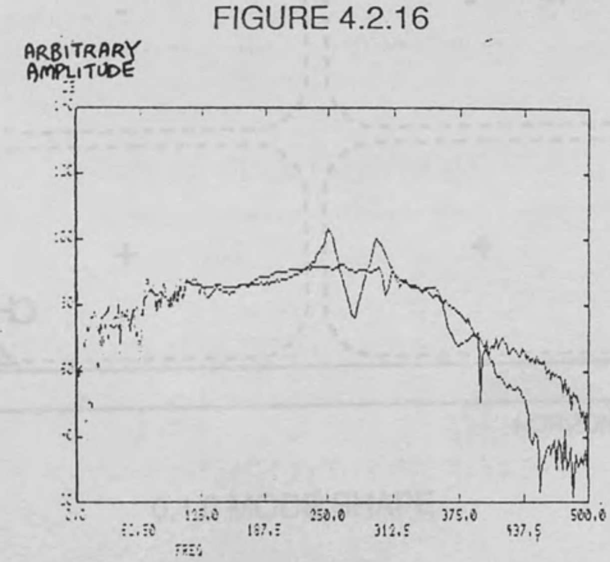
THEORETICAL ATTENUATION AT THE MONITOR

FIGURE 4.2.15



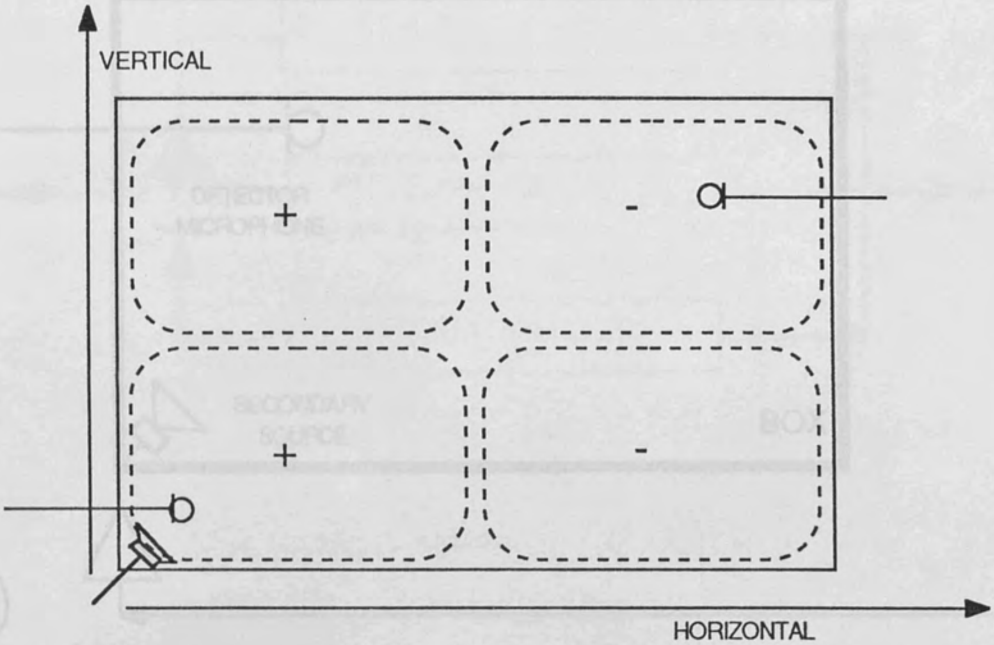
256 POINT FEEDFORWARD FILTER

FIGURE 4.2.16



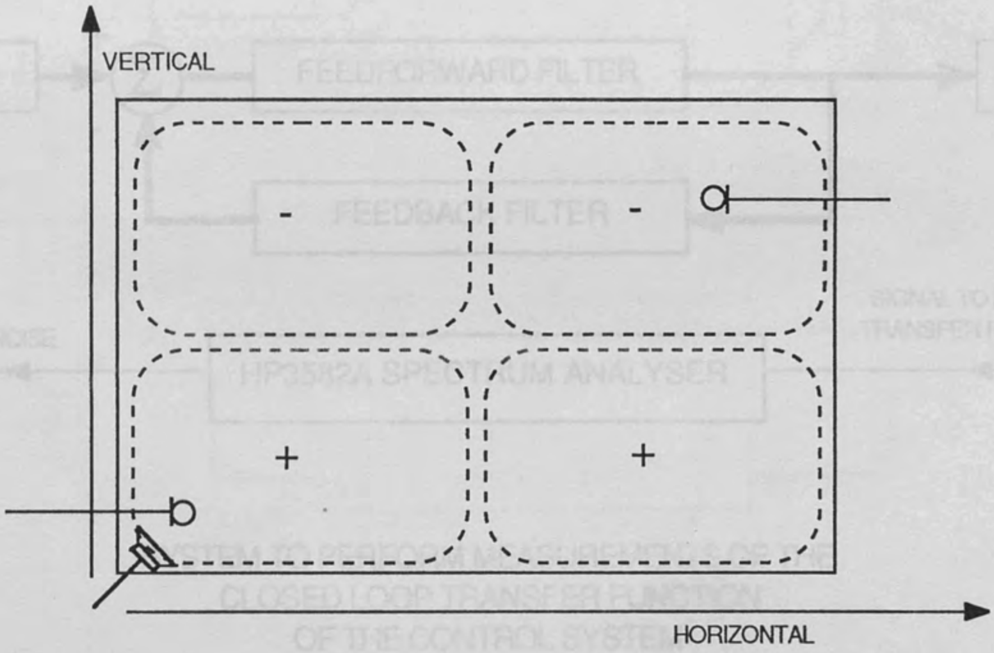
THEORETICAL ATTENUATION PRODUCED BY A 256 POINT FILTER

FIGURE 4.2.17



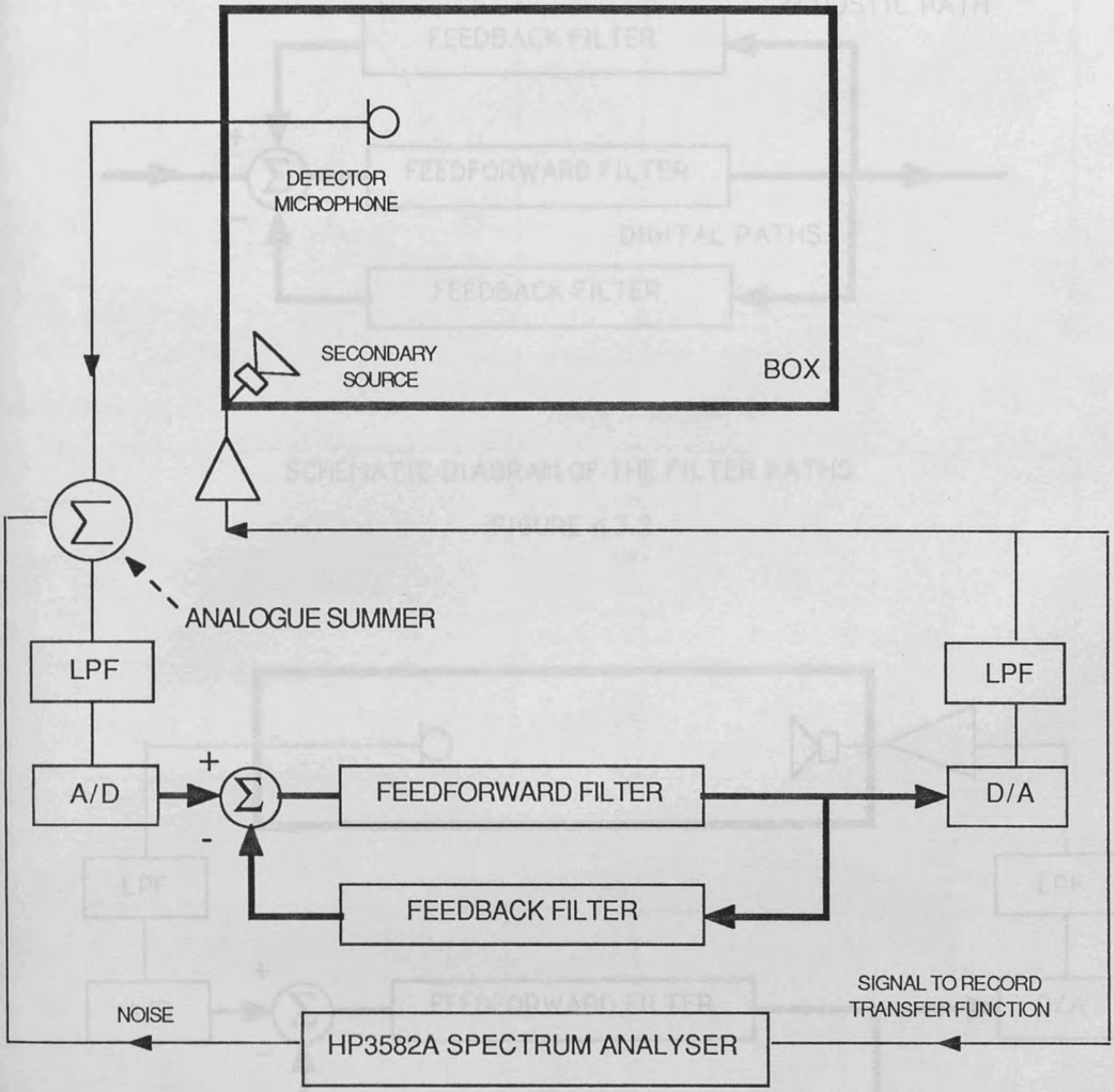
1,0,0 MODE SHAPE

FIGURE 4.2.18



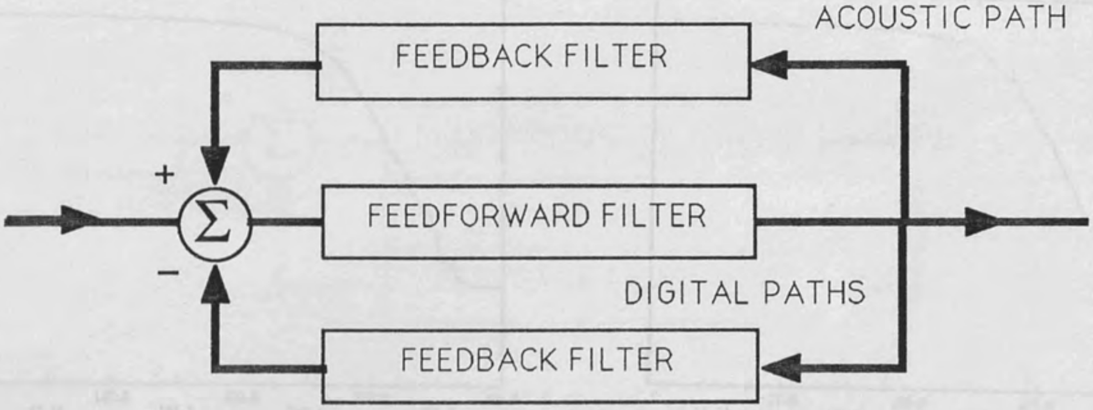
0,1,0 MODE SHAPE

FIGURE 4.2.19



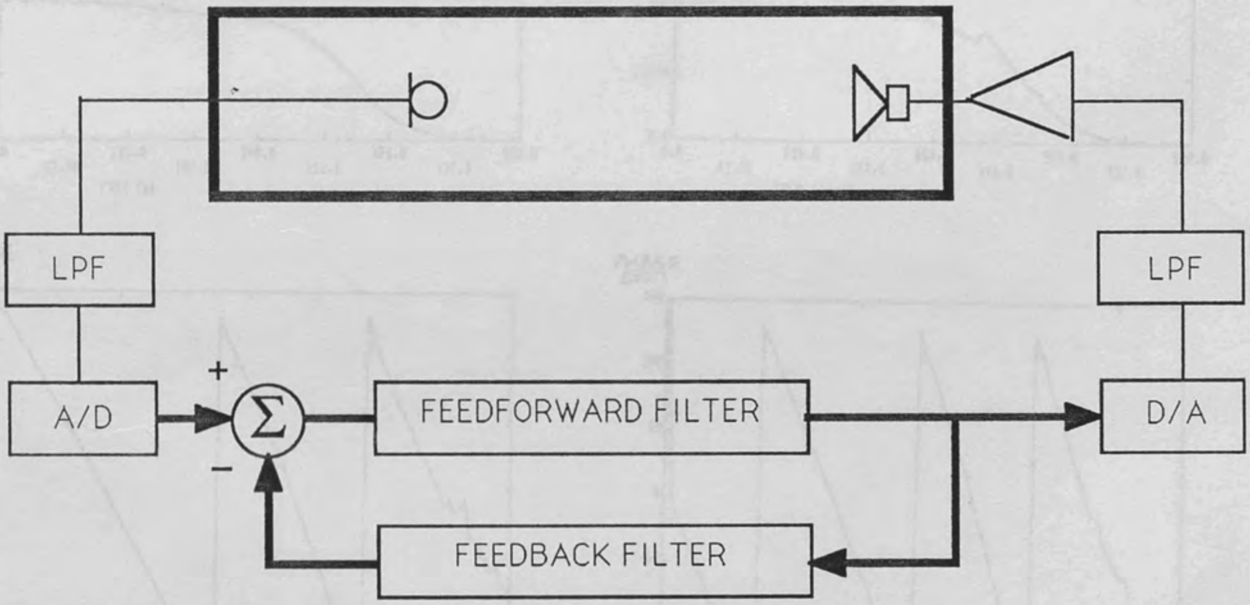
SYSTEM TO PERFORM MEASUREMENTS OF THE
CLOSED LOOP TRANSFER FUNCTION
OF THE CONTROL SYSTEM
OPERATING ON THE ENCLOSURE

FIGURE 4.3.1



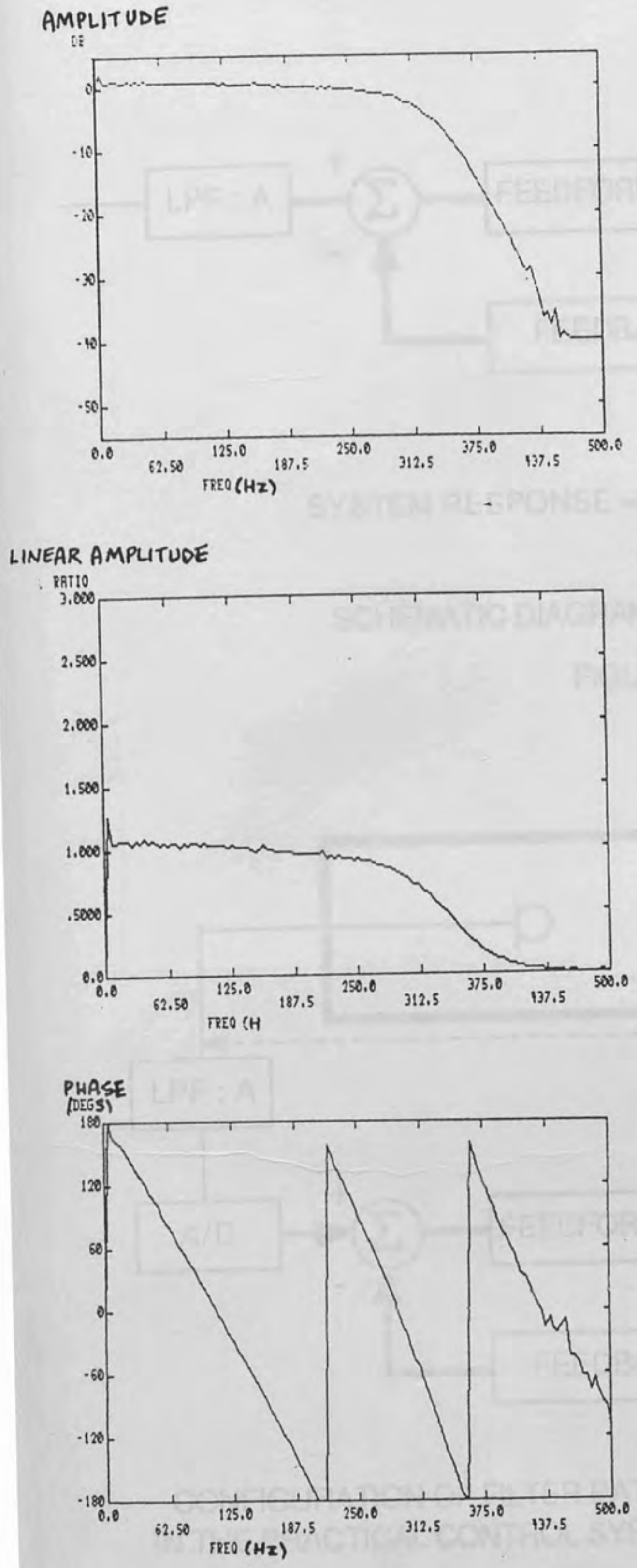
SCHEMATIC DIAGRAM OF THE FILTER PATHS

FIGURE 4.3.2



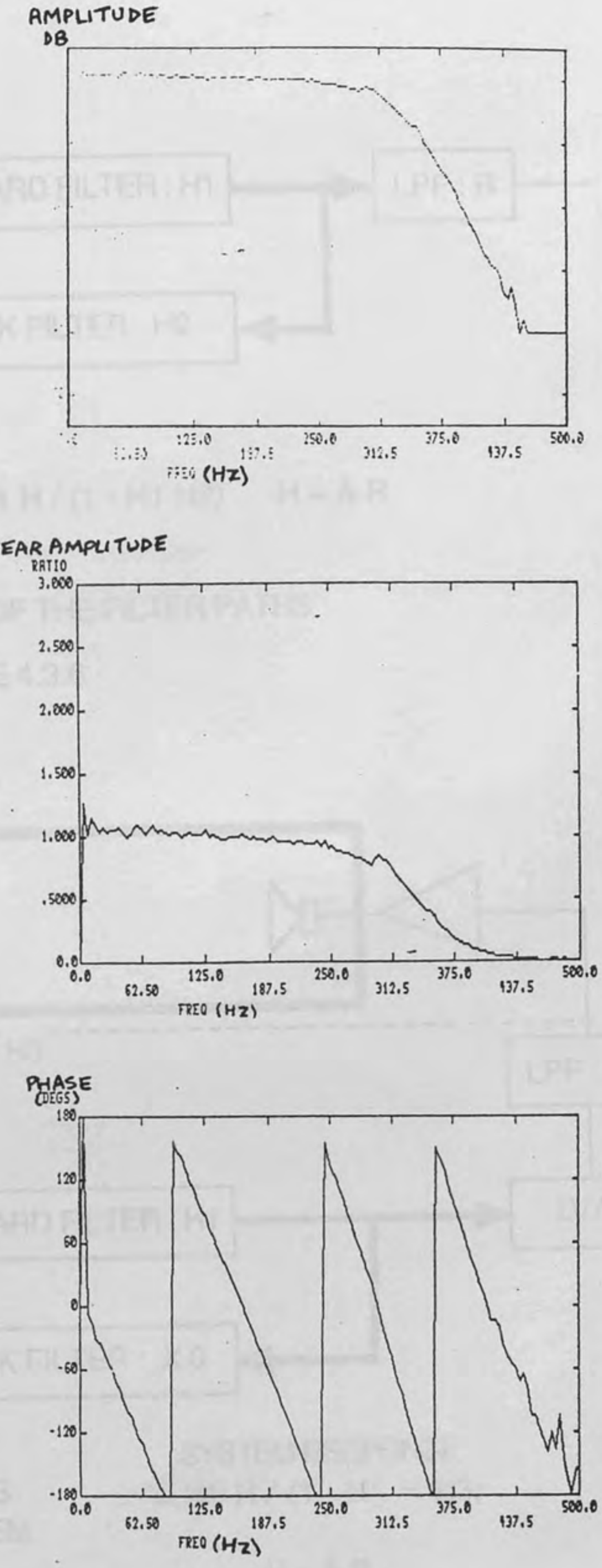
CONFIGURATION OF FILTER PATHS
IN THE PRACTICAL CONTROL SYSTEM

FIGURE 4.3.3



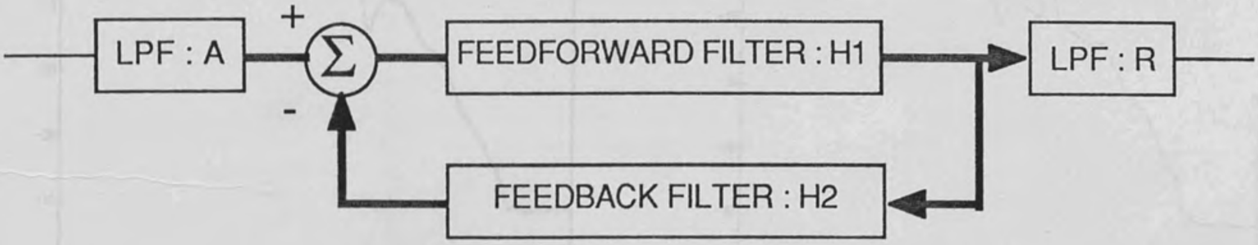
CLOSED LOOP TRANSFER FUNCTION
OF LOW PASS FILTERS AND ONE
CYCLE DELAY THROUGH THE DIGITAL SYSTEM

FIGURE 4.3.4



CLOSED LOOP TRANSFER FUNCTION
THE CONTROL SYSTEM WITH BOTH FEEDBACK PATHS
CONNECTED AND A SINGLE CYCLE DELAY
THROUGH THE FEEDFORWARD FILTER

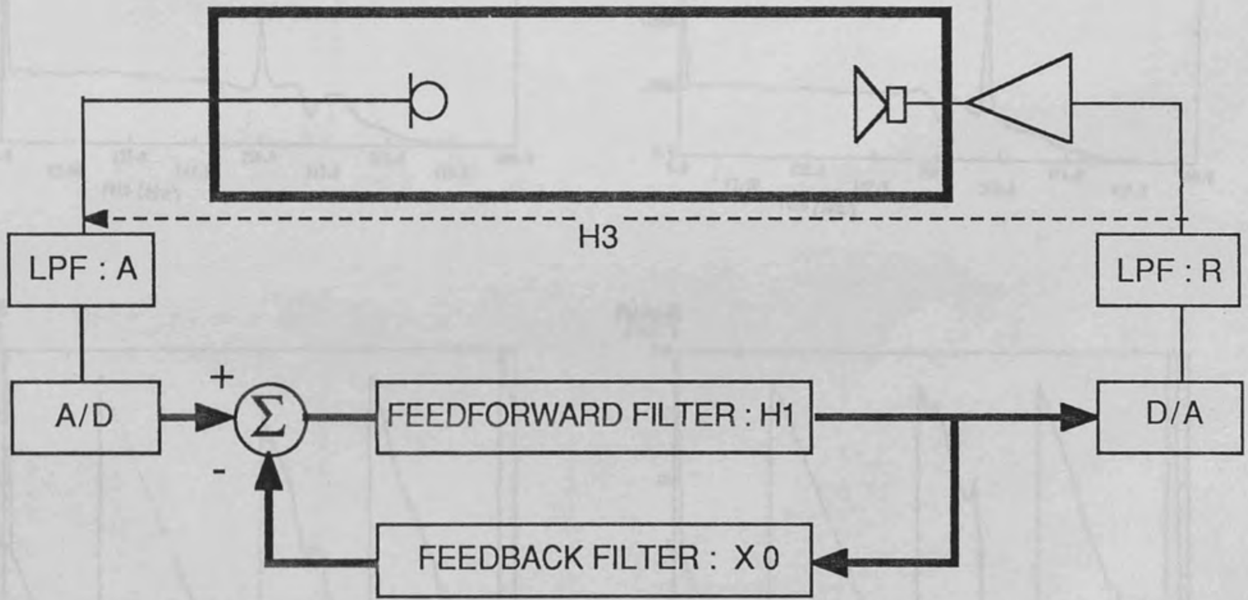
FIGURE 4.3.5



SYSTEM RESPONSE = H1 H / (1 - H1 H2)   H = A R

SCHEMATIC DIAGRAM OF THE FILTER PATHS

FIGURE 4.3.6

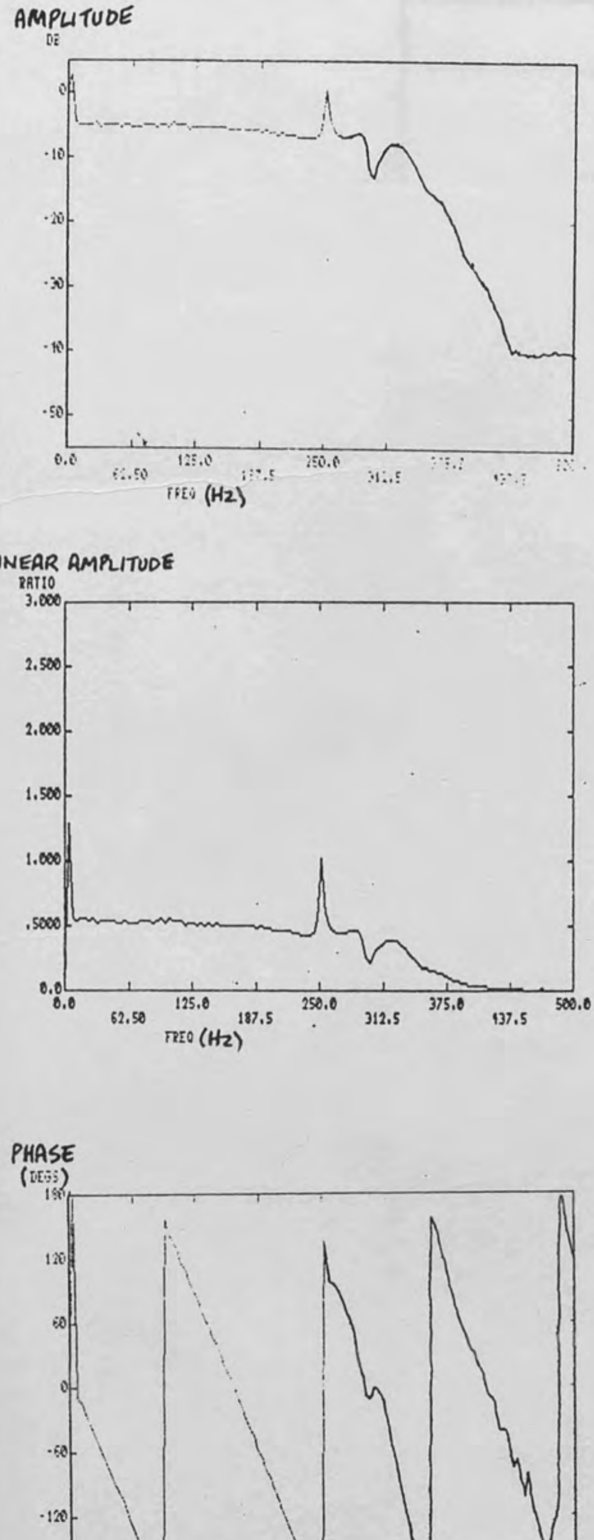


CONFIGURATION OF FILTER PATHS
IN THE PRACTICAL CONTROL SYSTEM

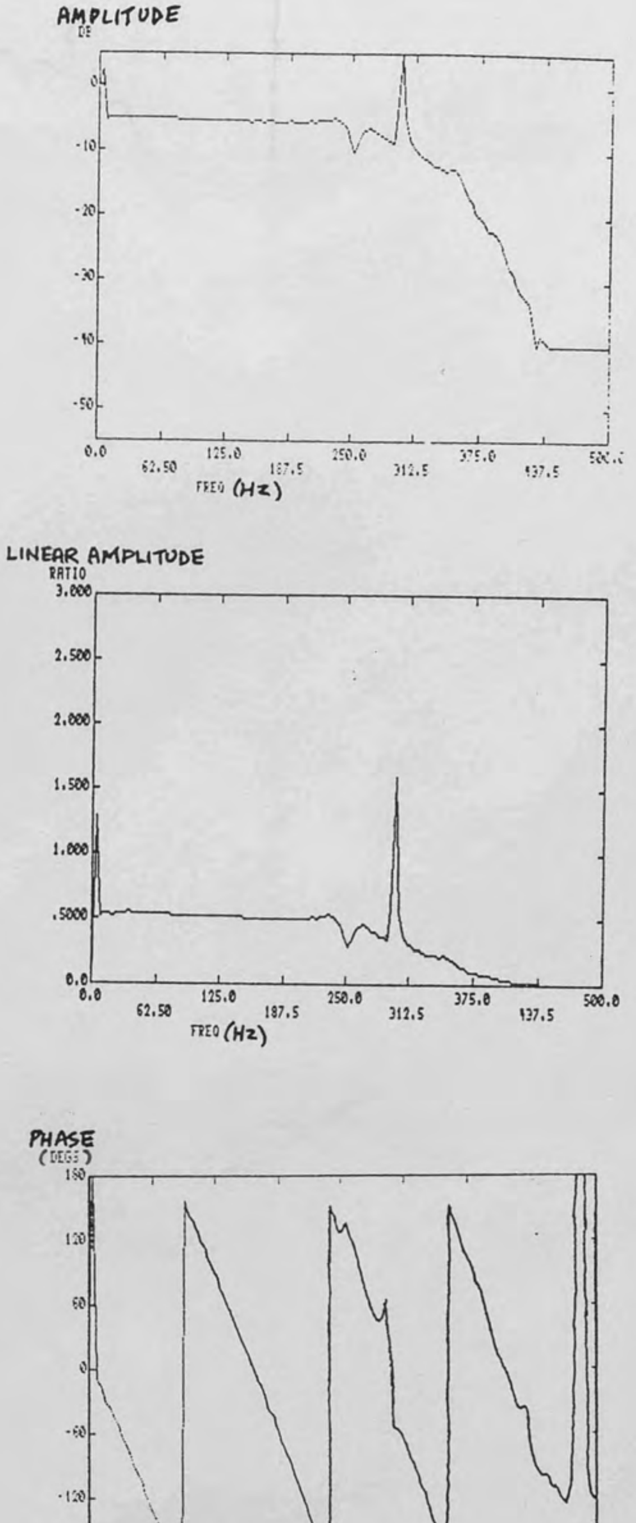
SYSTEM RESPONSE
= H1 H / (1 - H1 H H3)

H = A R

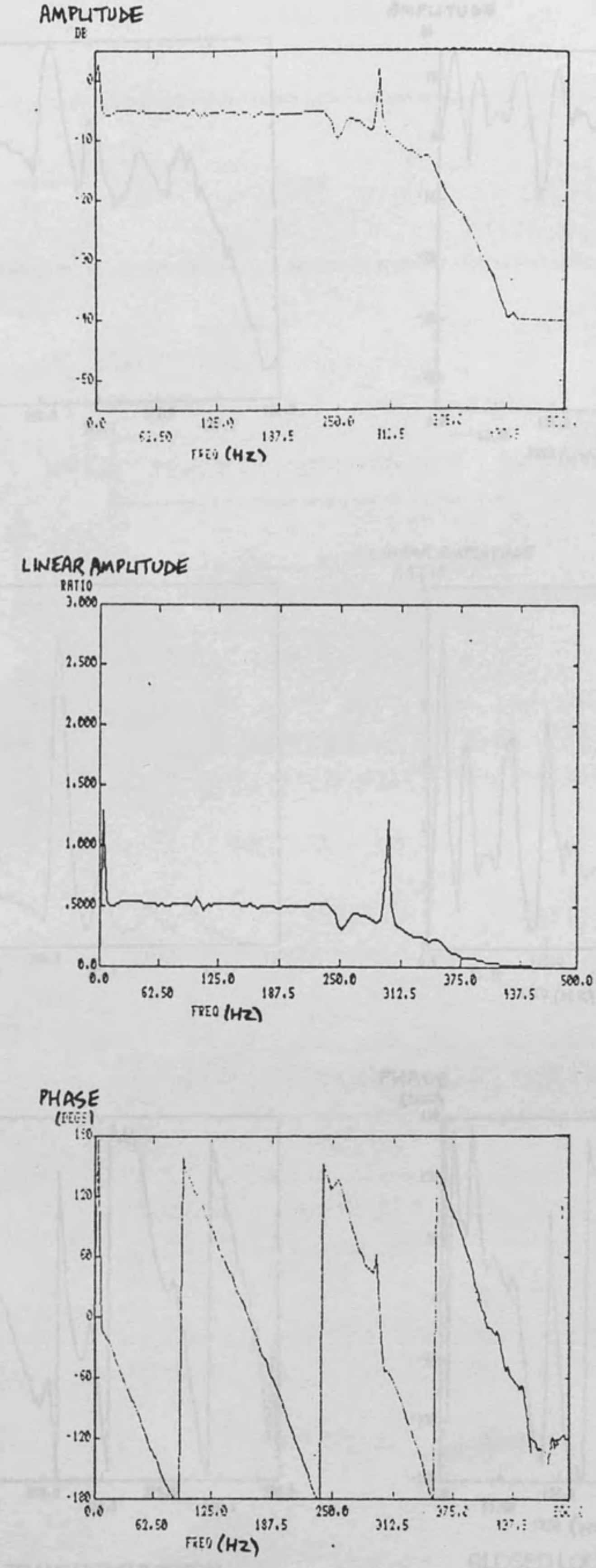
FIGURE 4.3.7



TRANSFER FUNCTION
OF FIGURE 4.3.6
FIGURE 4.3.8

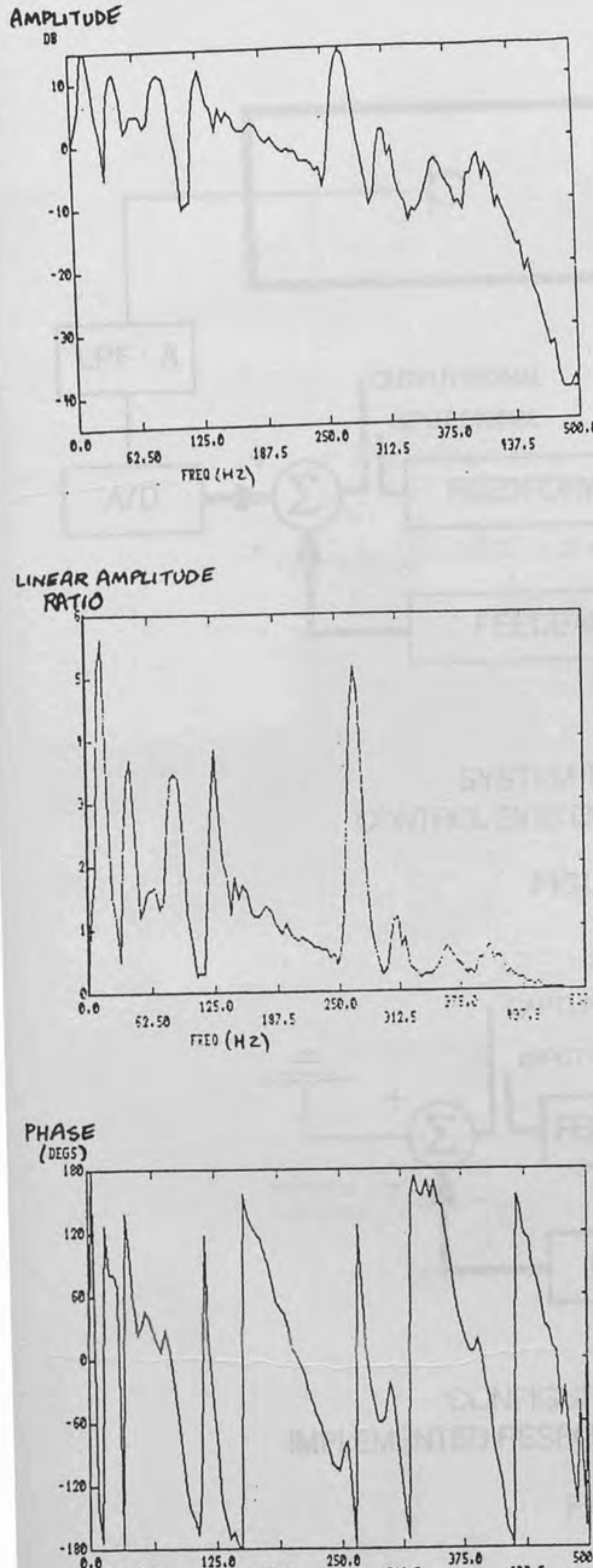


TRANSFER FUNCTION
OF FIGURE 4.6.7
FIGURE 4.3.9



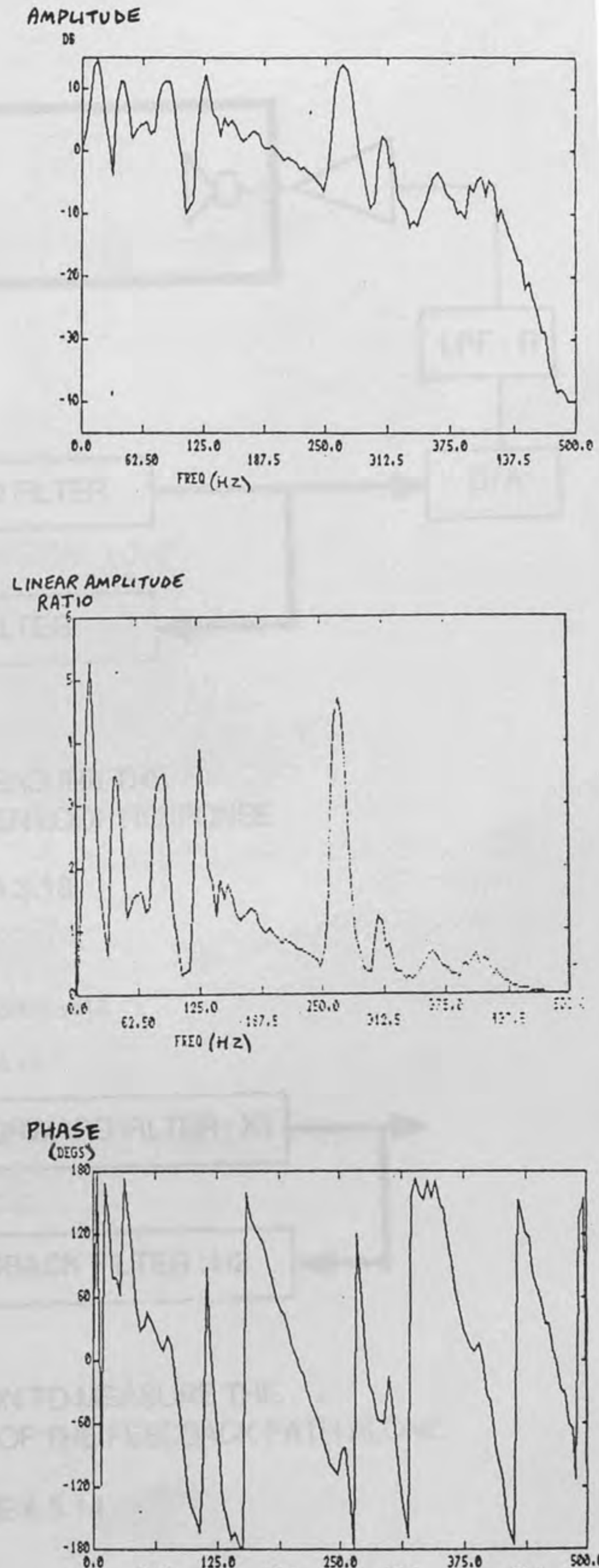
TRANSFER FUNCTION OF FIGURE 4.3.7 WITH THE LOUDSPEAKER TERMINALS REVERSED

FIGURE 4.3.10



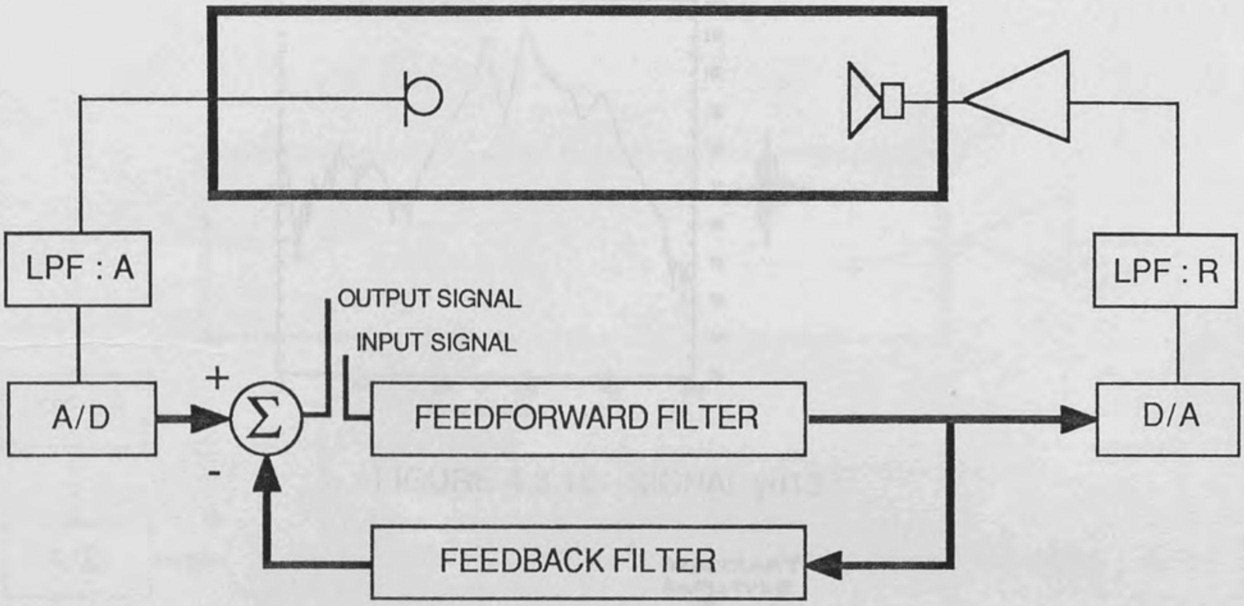
CLOSED LOOP TRANSFER FUNCTION
OF THE CONTROL SYSTEM WITH
THE FEEDBACK FILTERS DISCONNECTED

FIGURE 4.3.11



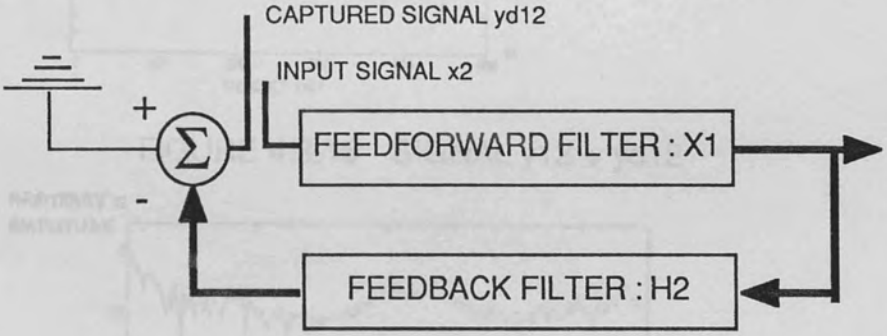
CLOSED LOOP TRANSFER FUNCTION
OF THE CONTROL SYSTEM WITH
BOTH FEEDBACK PATHS AND THE
FEEDFORWARD FILTER CONNECTED

FIGURE 4.3.12



SYSTEM TO MEASURE THE
CONTROL SYSTEM OPEN LOOP RESPONSE

FIGURE 4.3.13



CONFIGURATION TO MEASURE THE
IMPLEMENTED RESPONSE OF THE FEEDBACK PATH ALONE

FIGURE 4.3.14

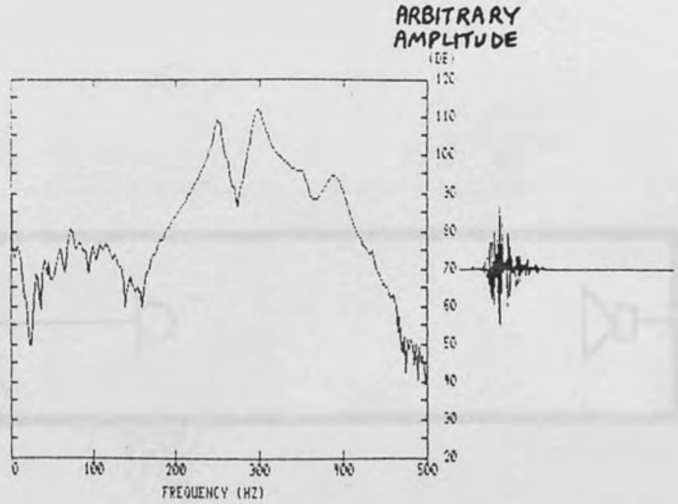


FIGURE 4.3.15 SIGNAL yd12

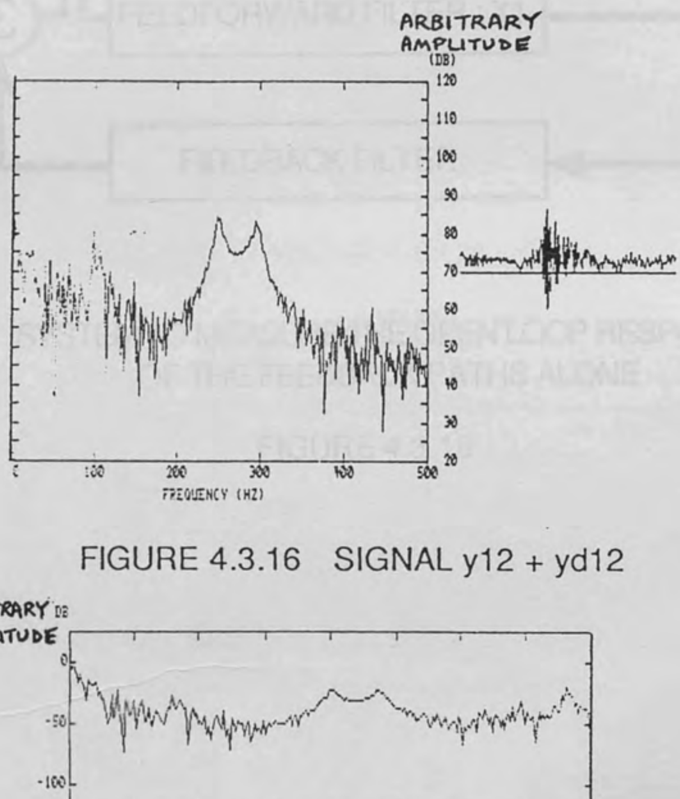
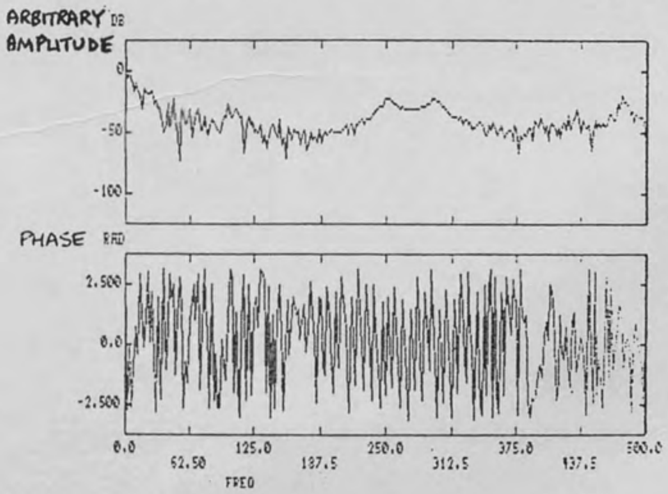
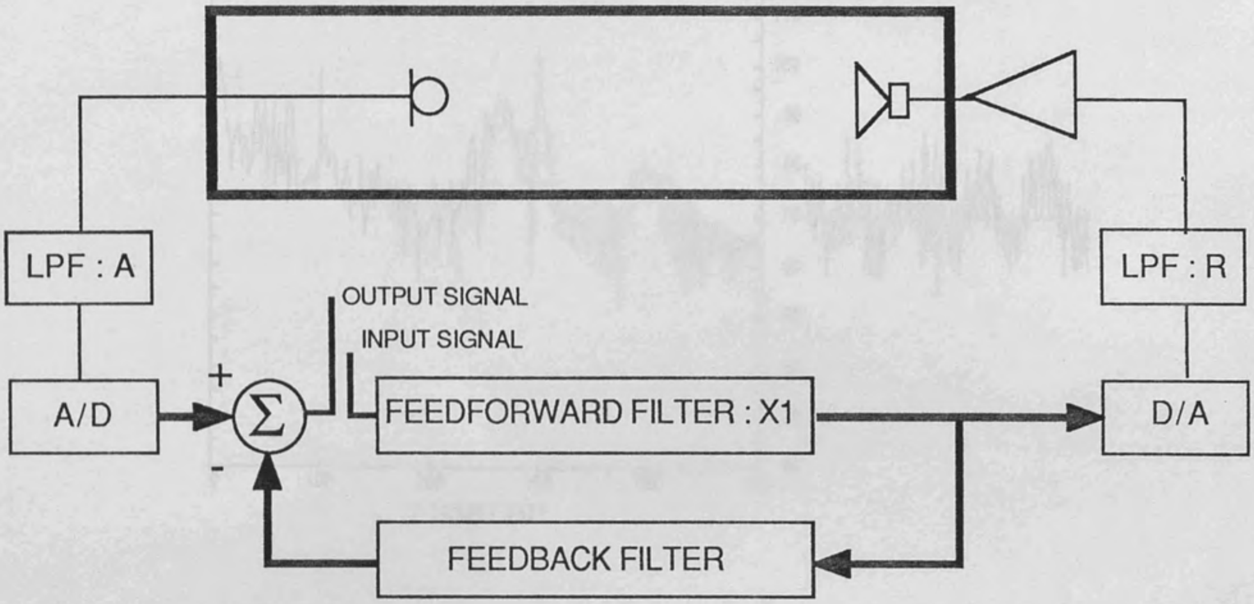


FIGURE 4.3.16 SIGNAL y12 + yd12



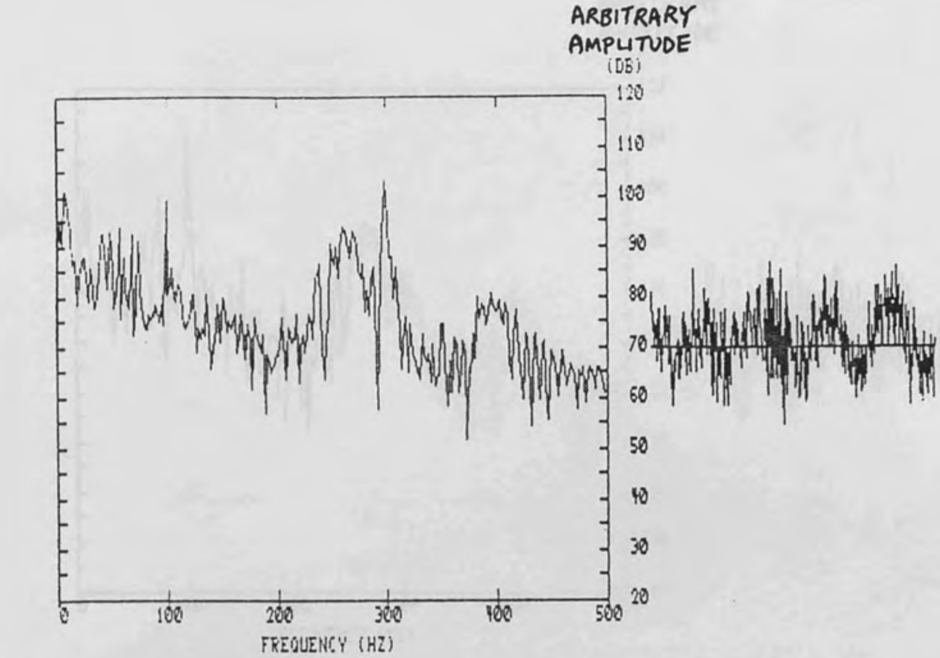
OPEN LOOP TRANSFER FUNCTION OF BOTH FEEDBACK PATHS AND
A SINGLE CYCLE DELAY THROUGH THE FORWARD PATH; (Y12+YD12)/X2

FIGURE 4.3.17



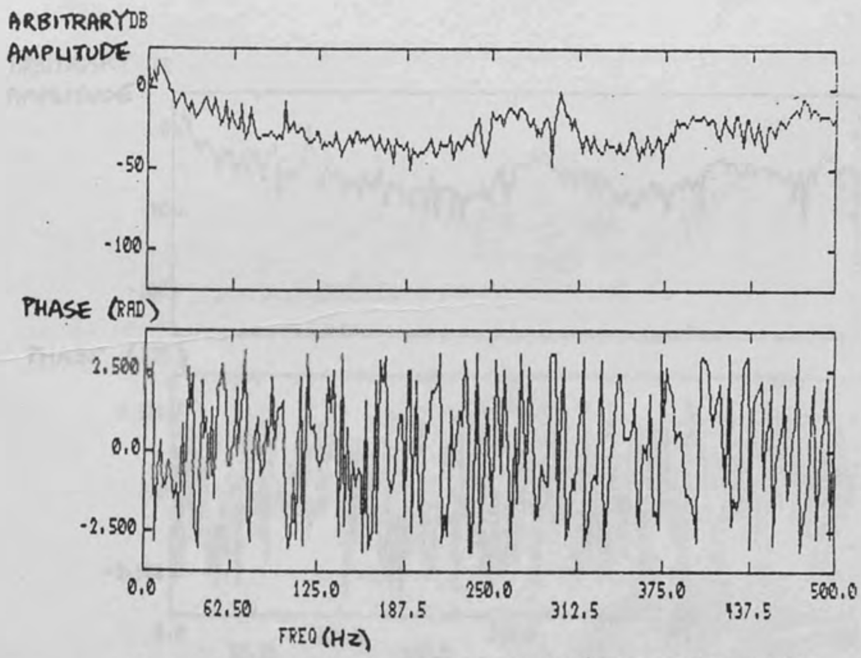
SYSTEM TO MEASURE THE OPEN LOOP RESPONSE
OF THE FEEDBACK PATHS ALONE

FIGURE 4.3.18



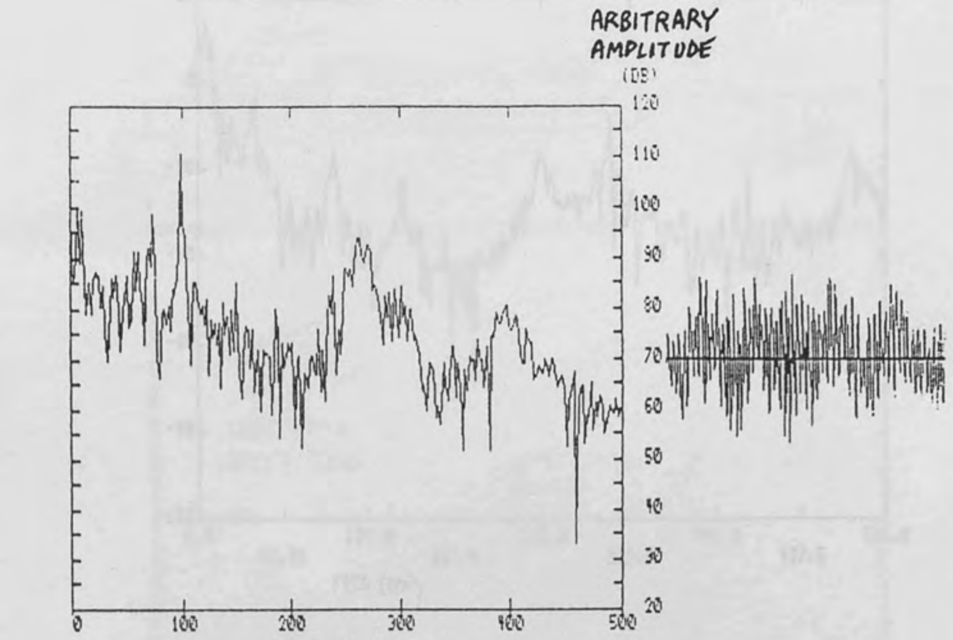
SIGNAL CAPTURED IN FIGURE 4.3.18

FIGURE 4.3.19



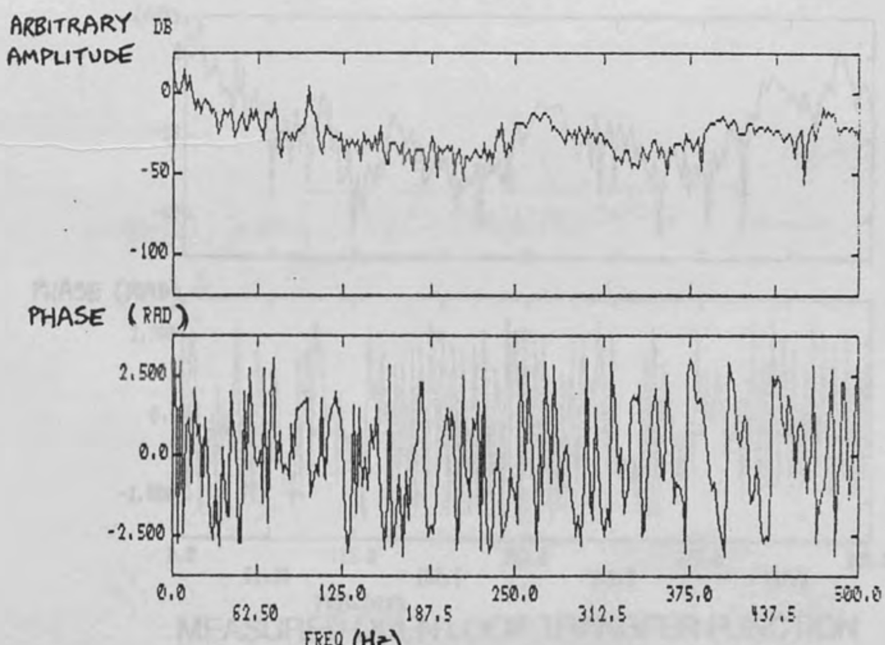
MEASURED OPEN LOOP TRANSFER FUNCTION
OF BOTH FEEDBACK PATHS
AND A SINGLE CYCLE DELAY THROUGH THE FORWARD PATH

FIGURE 4.3.20



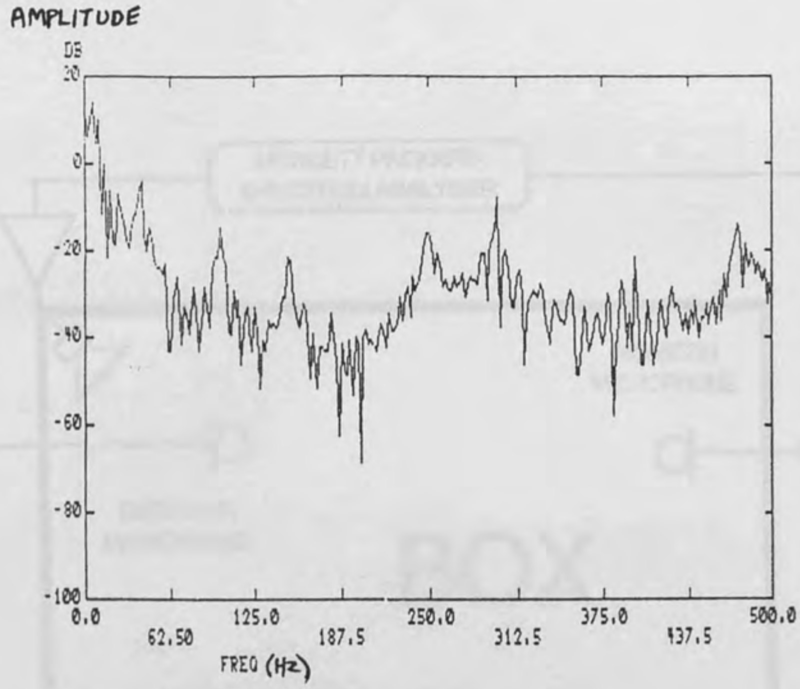
SIGNAL CAPTURED IN FIGURE 4.3.13

FIGURE 4.3.21

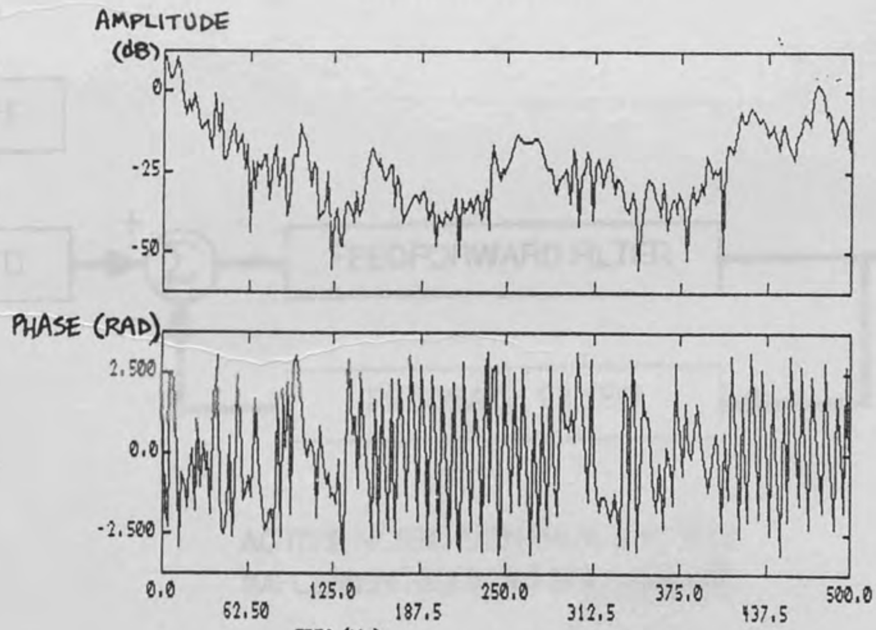


MEASURED OPEN LOOP TRANSFER FUNCTION
OF THE CONTROL SYSTEM

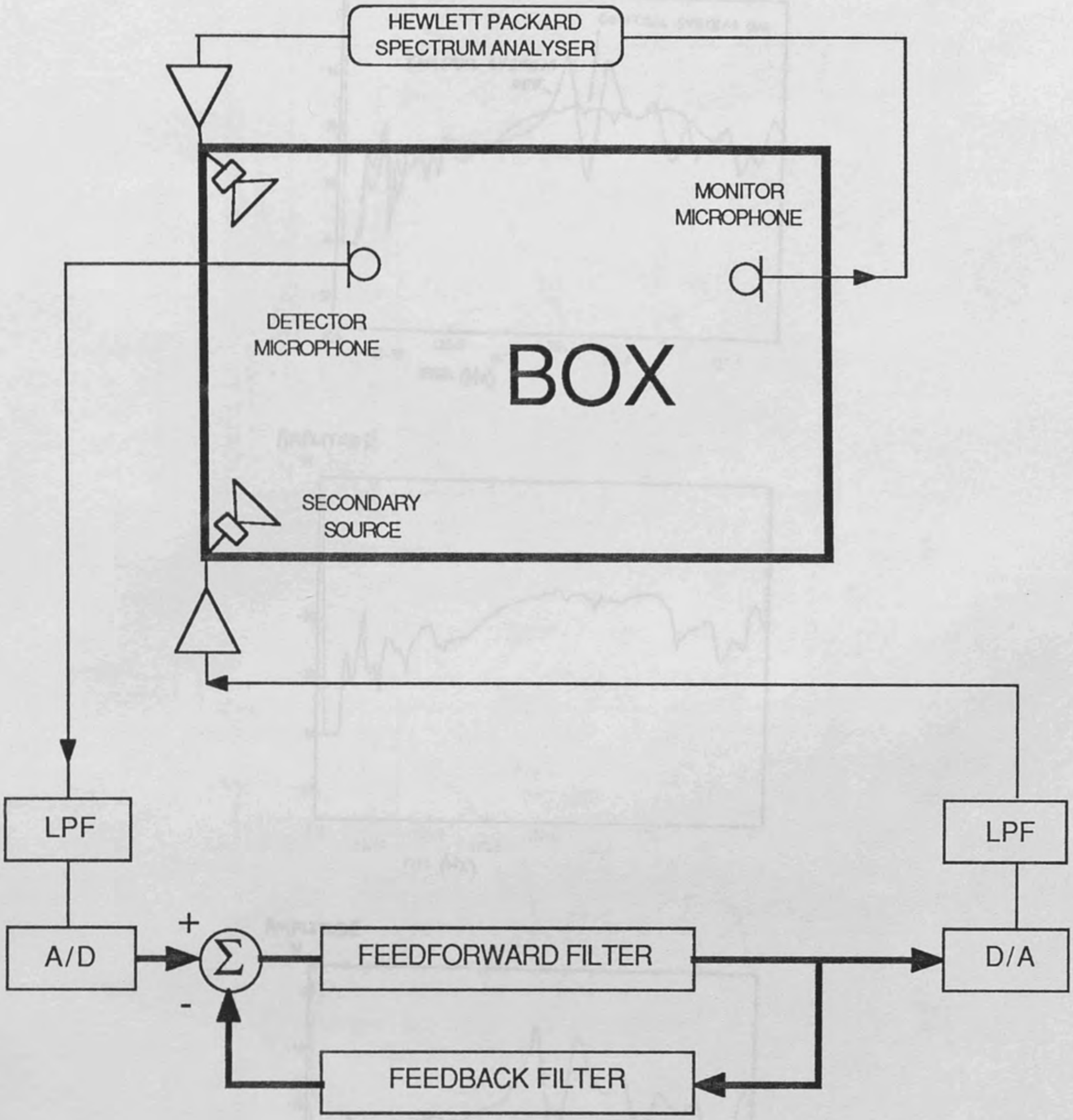
FIGURE 4.3.22



MEASURED OPEN LOOP TRANSFER FUNCTION
OF BOTH FEEDBACK PATHS
AND A SINGLE CYCLE DELAY THROUGH THE FORWARD PATH;
SYSTEM RESPONSE DERIVED WITH A TEST SIGNAL
STARTING AT ONLY 10 HZ
FIGURE 4.3.23

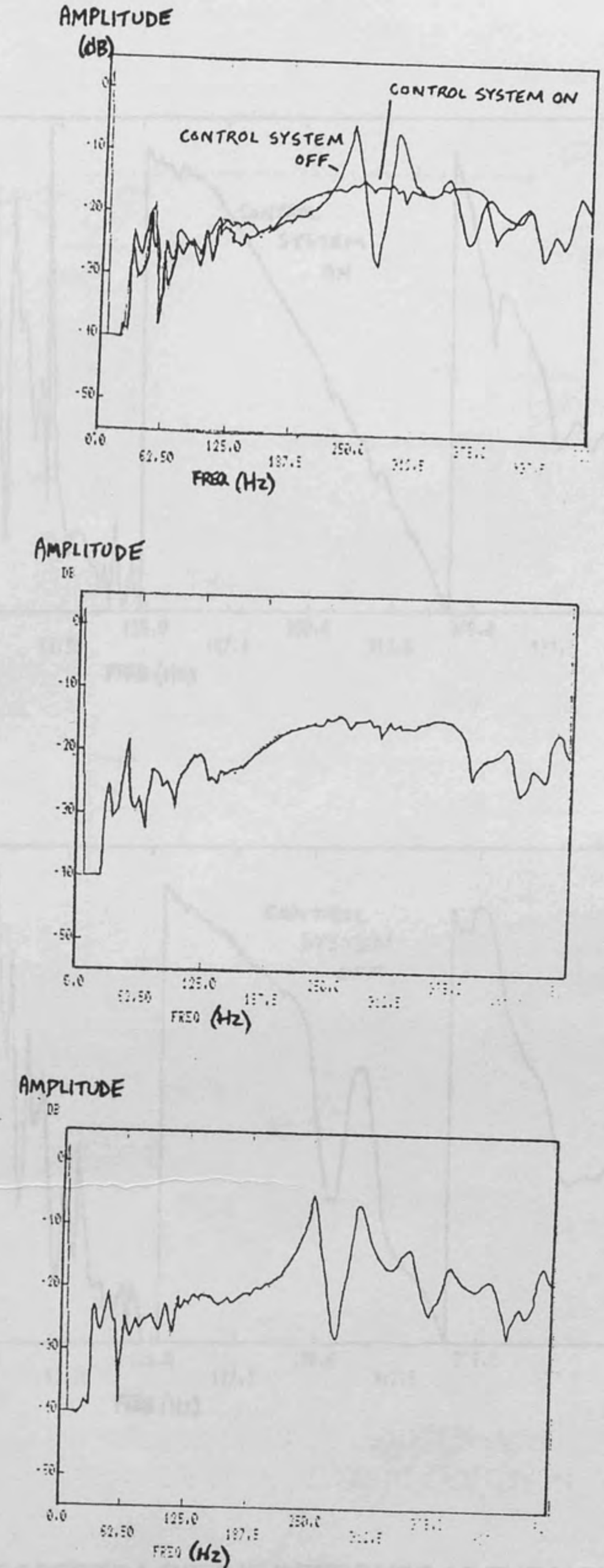


MEASURED OPEN LOOP TRANSFER FUNCTION
OF THE CONTROL SYSTEM;
SYSTEM RESPONSE DERIVED WITH A TEST SIGNAL
STARTING AT ONLY 10 HZ
FIGURE 4.3.24



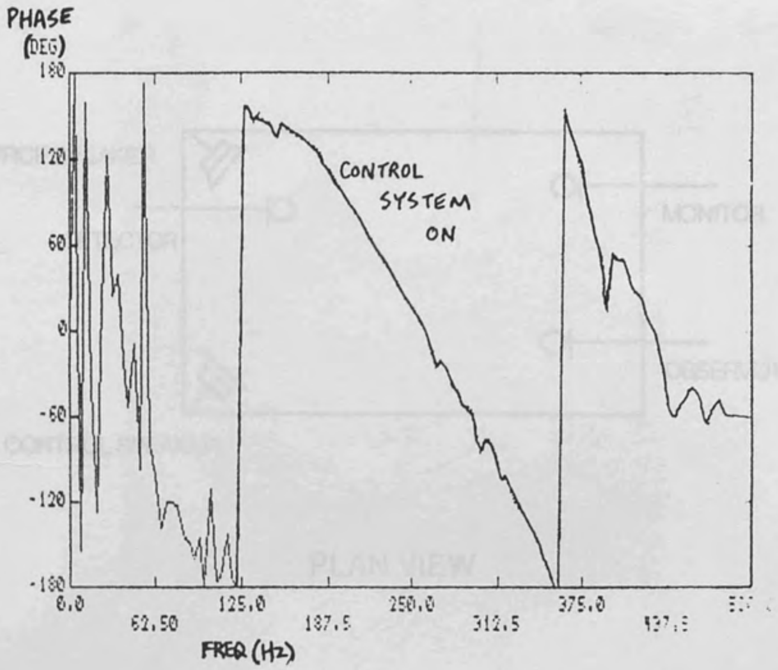
ACTIVE NOISE CONTROL SYSTEM
IMPLEMENTED IN AN ENCLOSURE

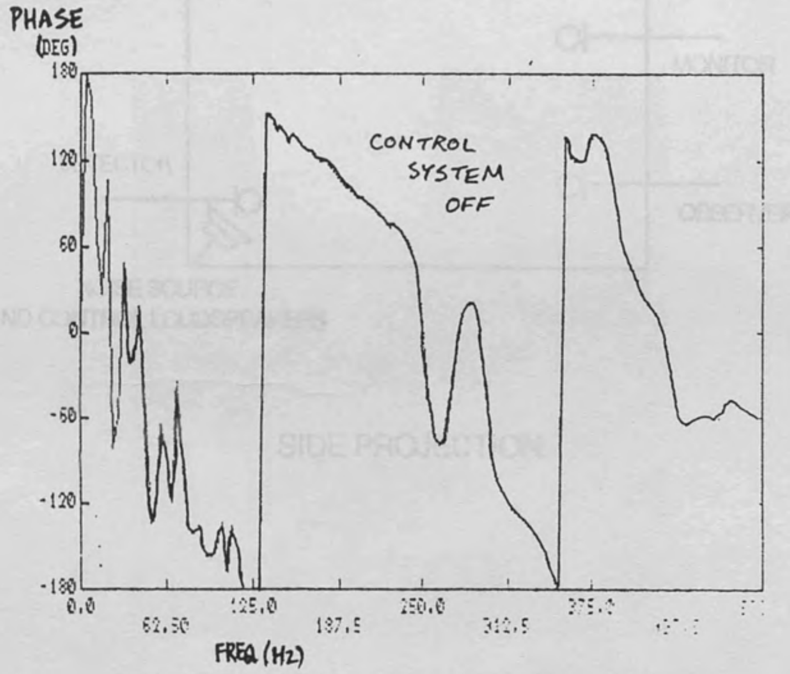
FIGURE 4.3.25



AMPLITUDE SPECTRA OF THE RESPONSE AT THE MONITOR MICROPHONE WITH AND WITHOUT THE CONTROL SYSTEM OPERATING

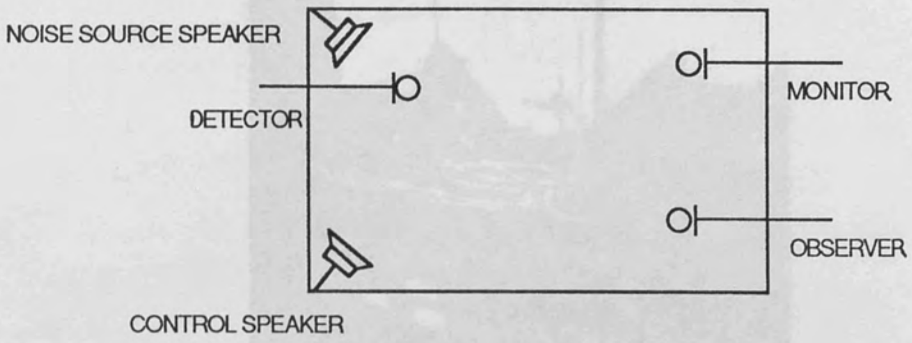
FIGURE 4.3.26



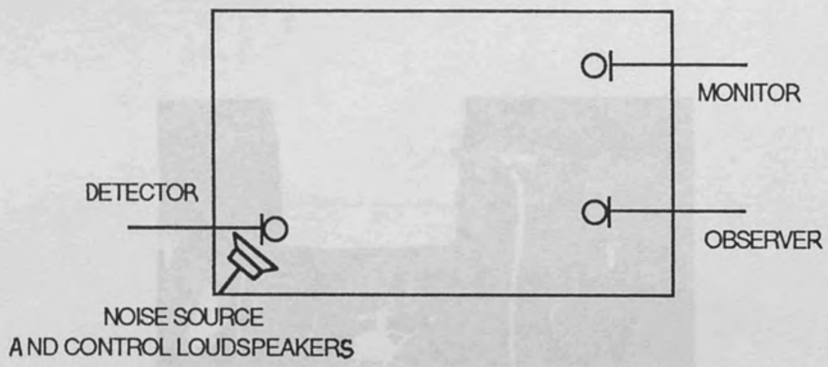


PHASE SPECTRA OF THE RESPONSE AT THE MONITOR
MICROPHONE WITH AND WITHOUT THE CONTROL SYSTEM OPERATING

FIGURE 4.3.27



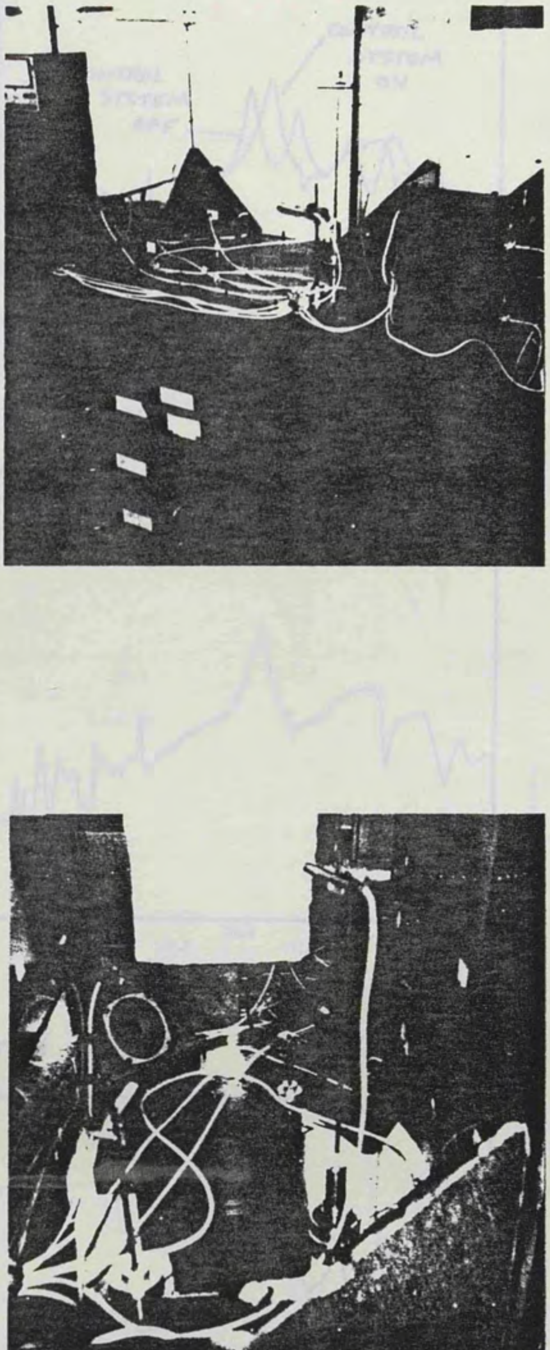
PLAN VIEW



SIDE PROJECTION

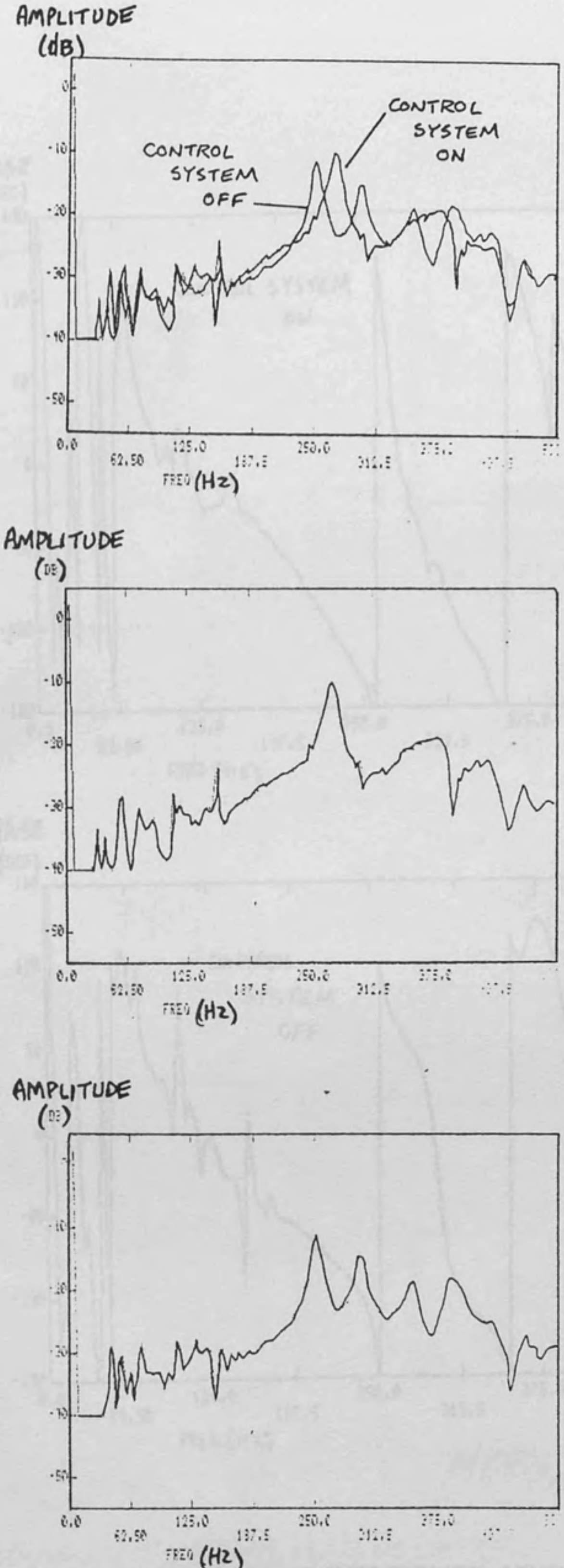
CONTROL SYSTEM TRANSDUCER POSITIONS
PLAN VIEW AND SIDE PROJECTION

FIGURE 4.3.28



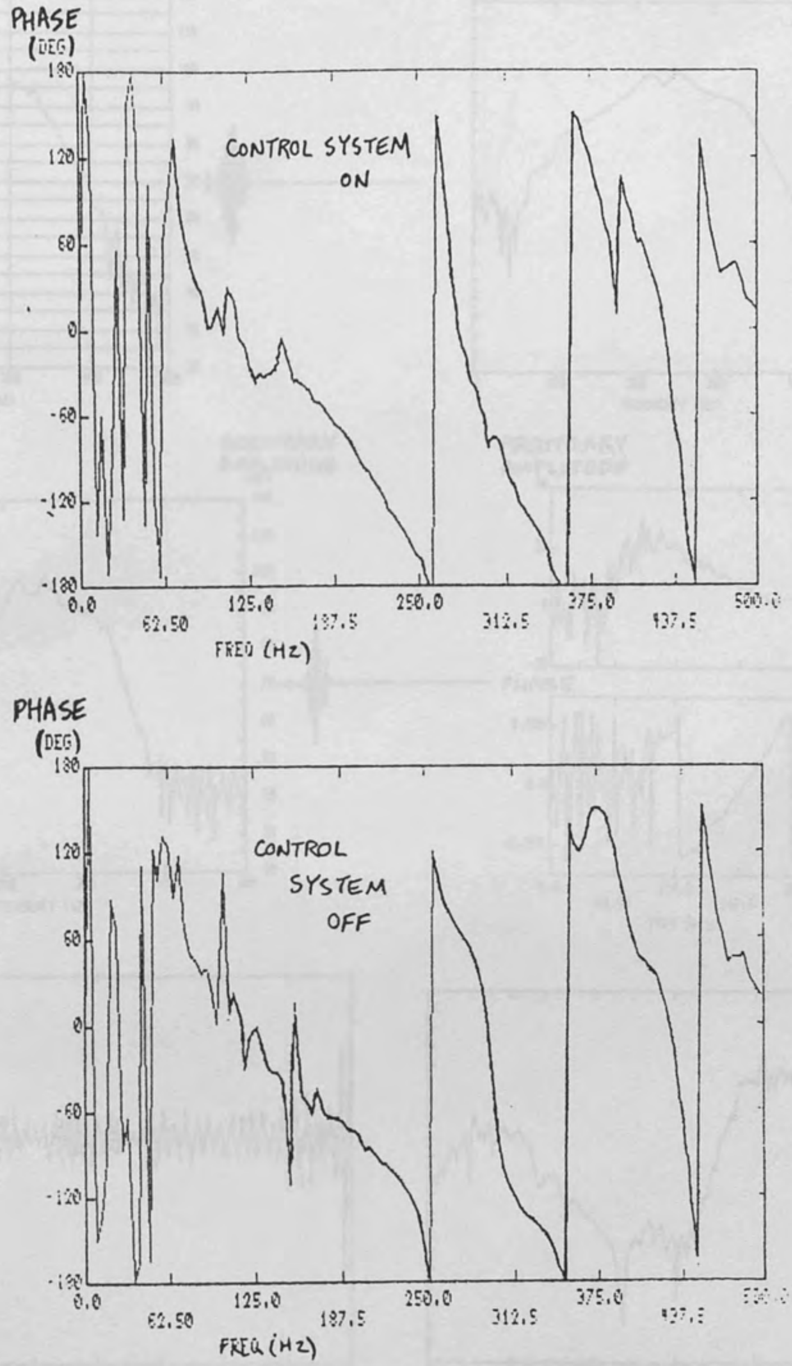
PICTURES SHOWING THE POSITIONS OF THE OF THE TRANSDUCERS
INSIDE THE ENCLOSURE

FIGURE 4.3.29



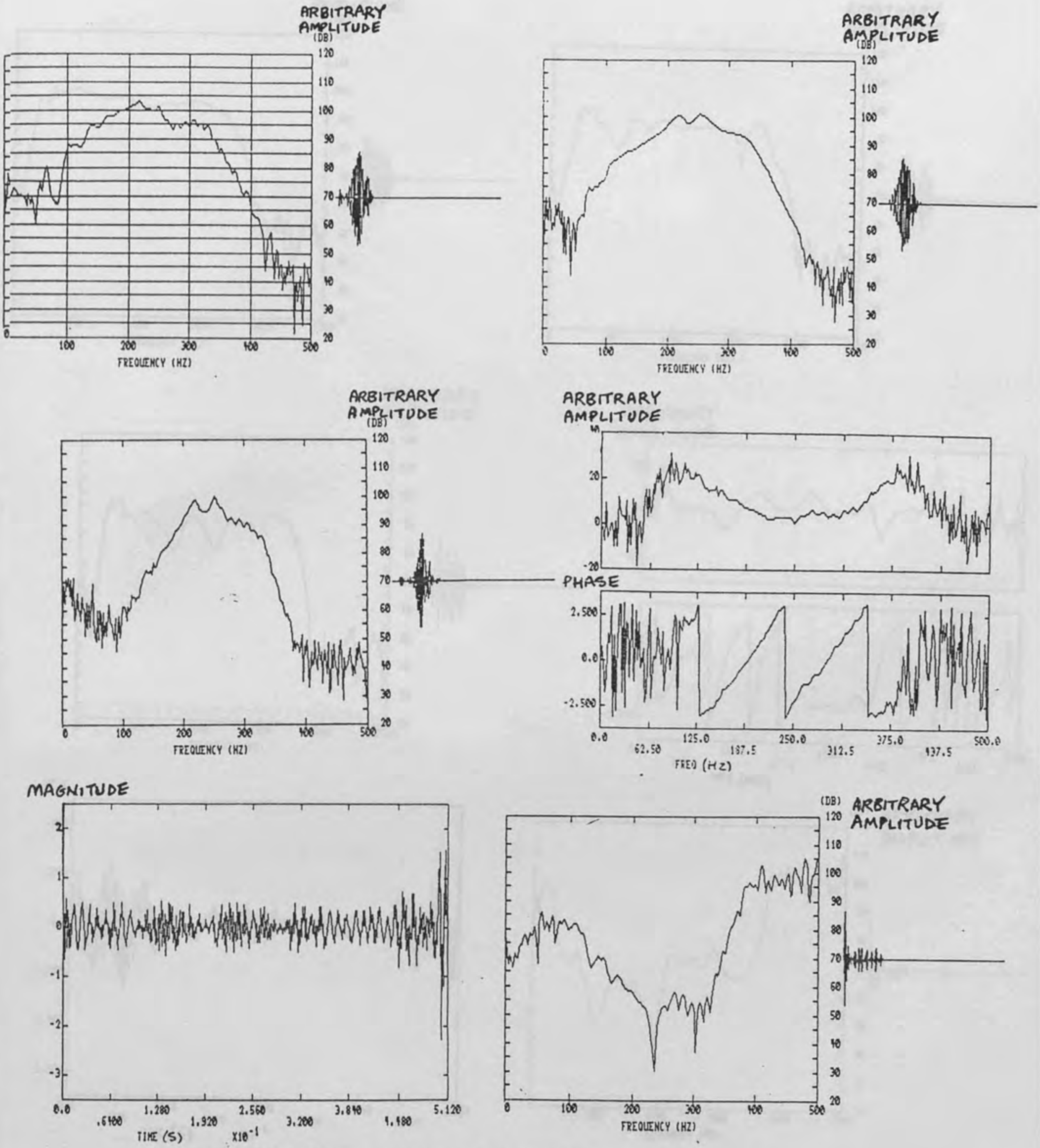
AMPLITUDE SPECTRA OF THE RESPONSE AT THE OBSERVER MICROPHONE WITH AND WITHOUT THE CONTROL SYSTEM OPERATING

FIGURE 4.3.30



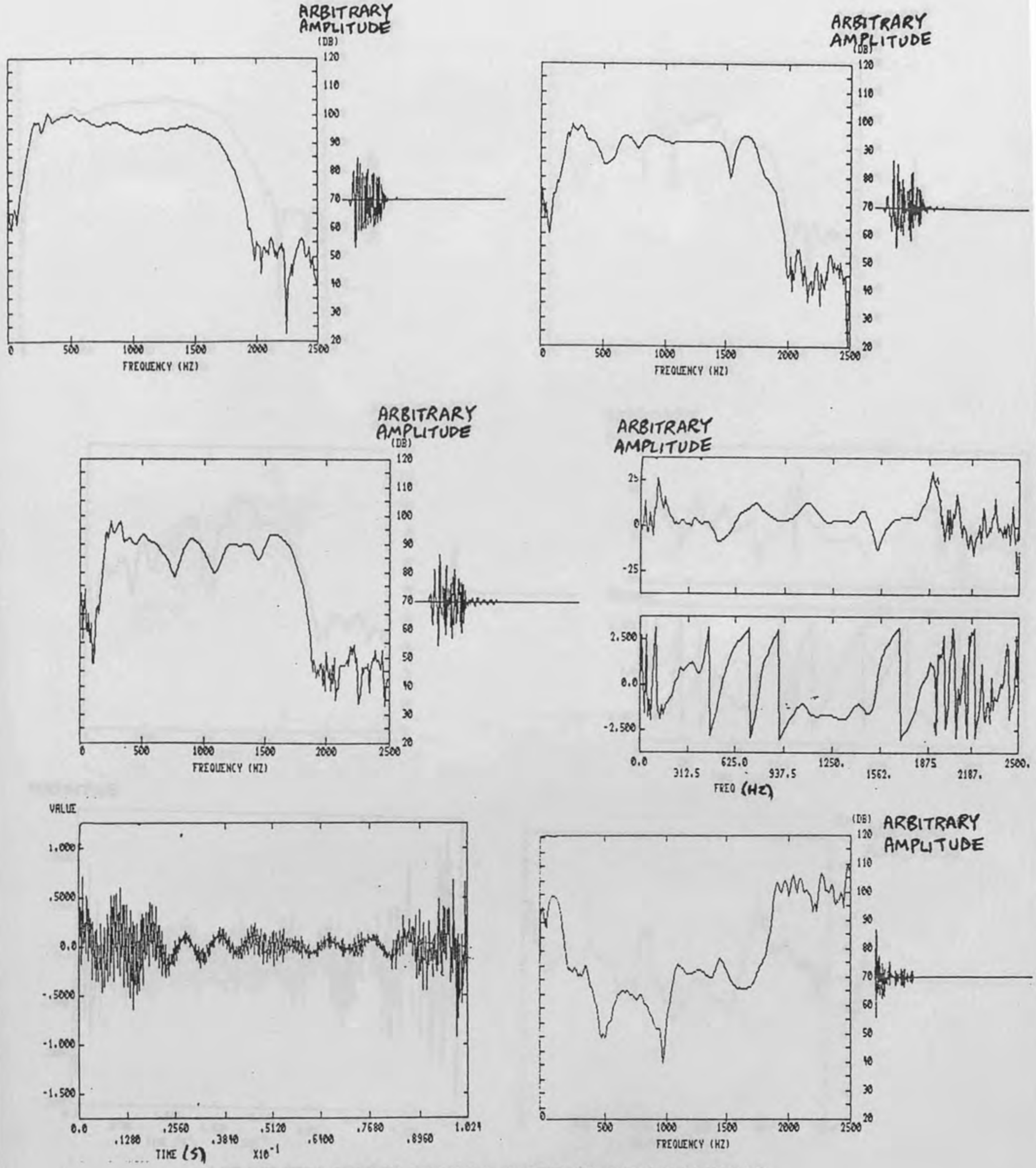
PHASE SPECTRA OF THE RESPONSE AT THE OBSERVER MICROPHONE WITH AND WITHOUT THE CONTROL SYSTEM OPERATING

FIGURE 4.3.31



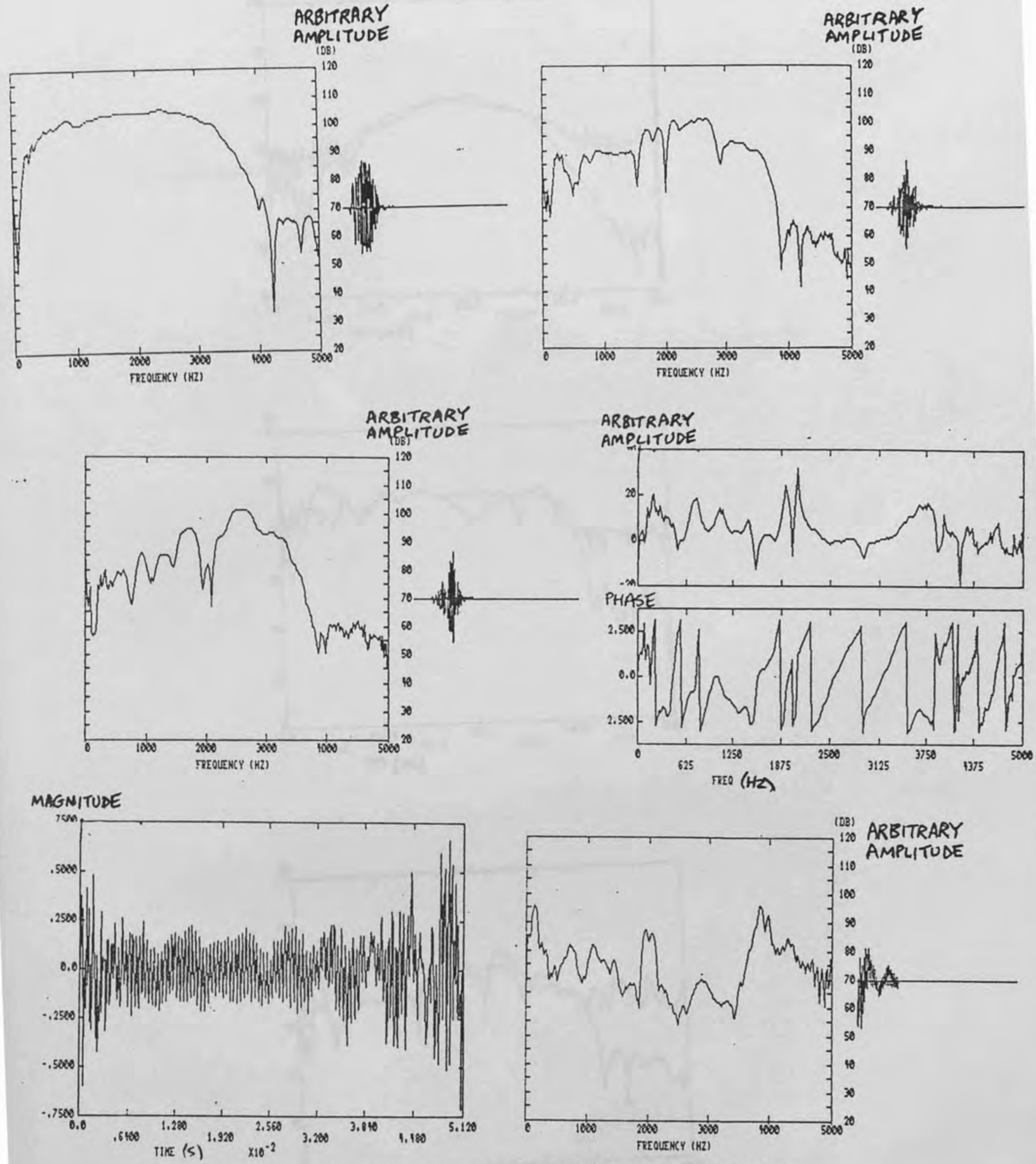
MEASUREMENTS AT 1KHz WITH HEAVY DAMPING;
SIGNALS y10, y30, y32,
AND THE SPECTRUM Y30/Y32 AND ITS INVERSE TRANSFORM
AND THE 128 POINT FIR FILTERDERIVED FROM THE MEASUREMENTS

FIGURE 4.4.1



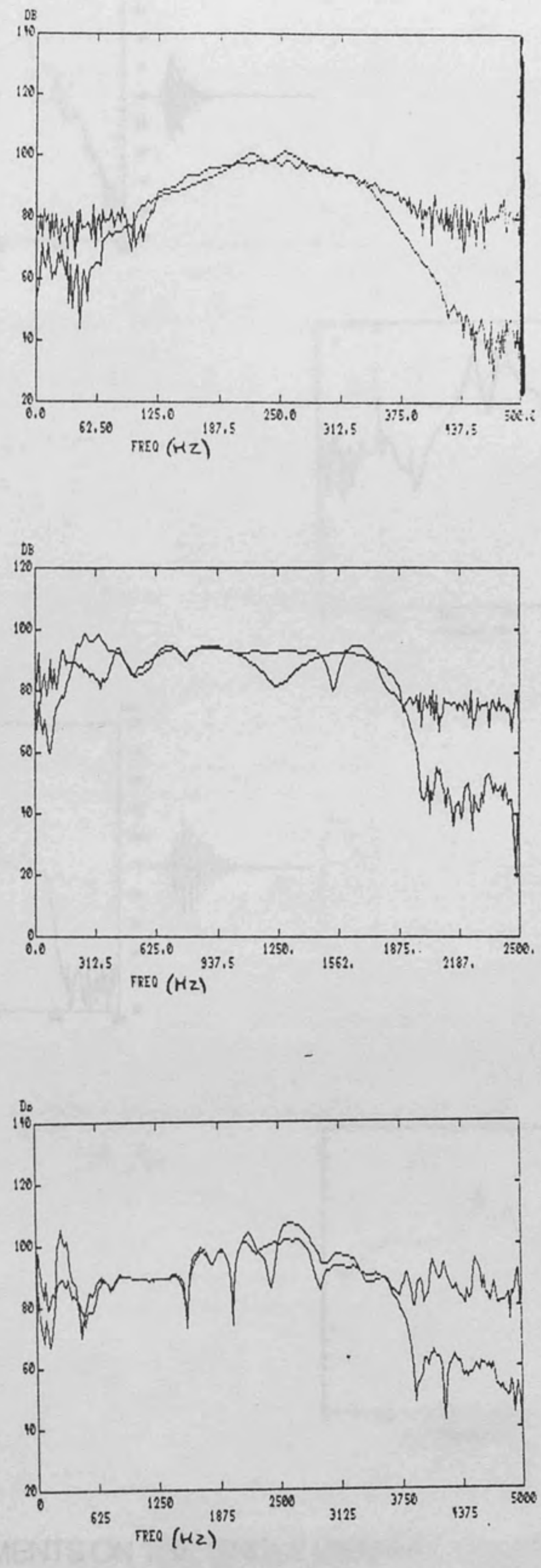
MEASUREMENTS AT 5KHz WITH HEAVY DAMPING;
SIGNALS y10, y30, y32,
AND THE SPECTRUM Y30/Y32 AND ITS INVERSE TRANSFORM
AND THE 128 POINT FIR FILTERDERIVED FROM THE MEASUREMENTS

FIGURE 4.4.2



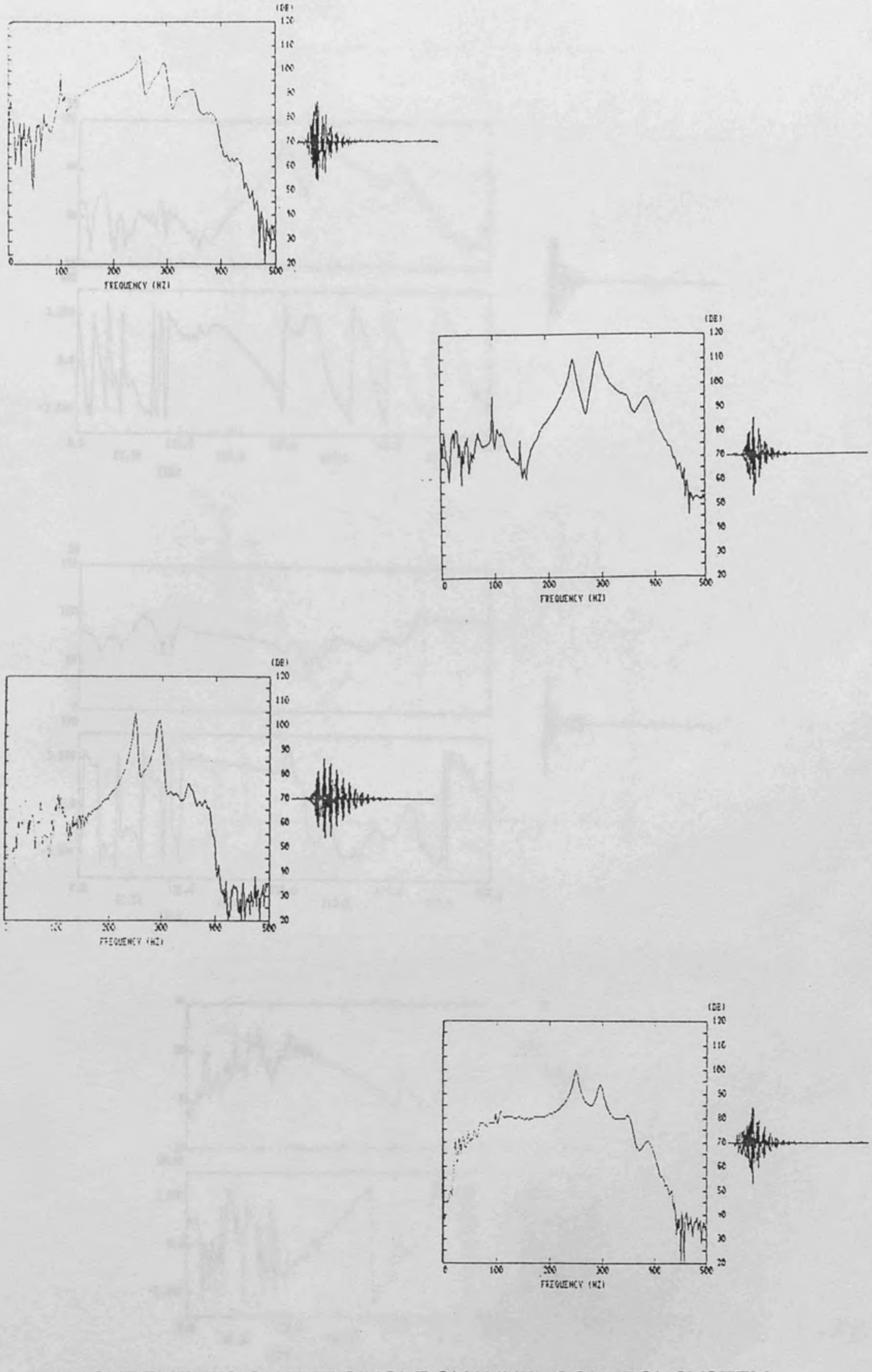
MEASUREMENTS AT 10KHz WITH HEAVY DAMPING;
SIGNALS y10, y30, y32,
AND THE SPECTRUM Y30/Y32 AND ITS INVERSE TRANSFORM
AND THE 128 POINT FIR FILTERDERIVED FROM THE MEASUREMENTS

FIGURE 4.4.3



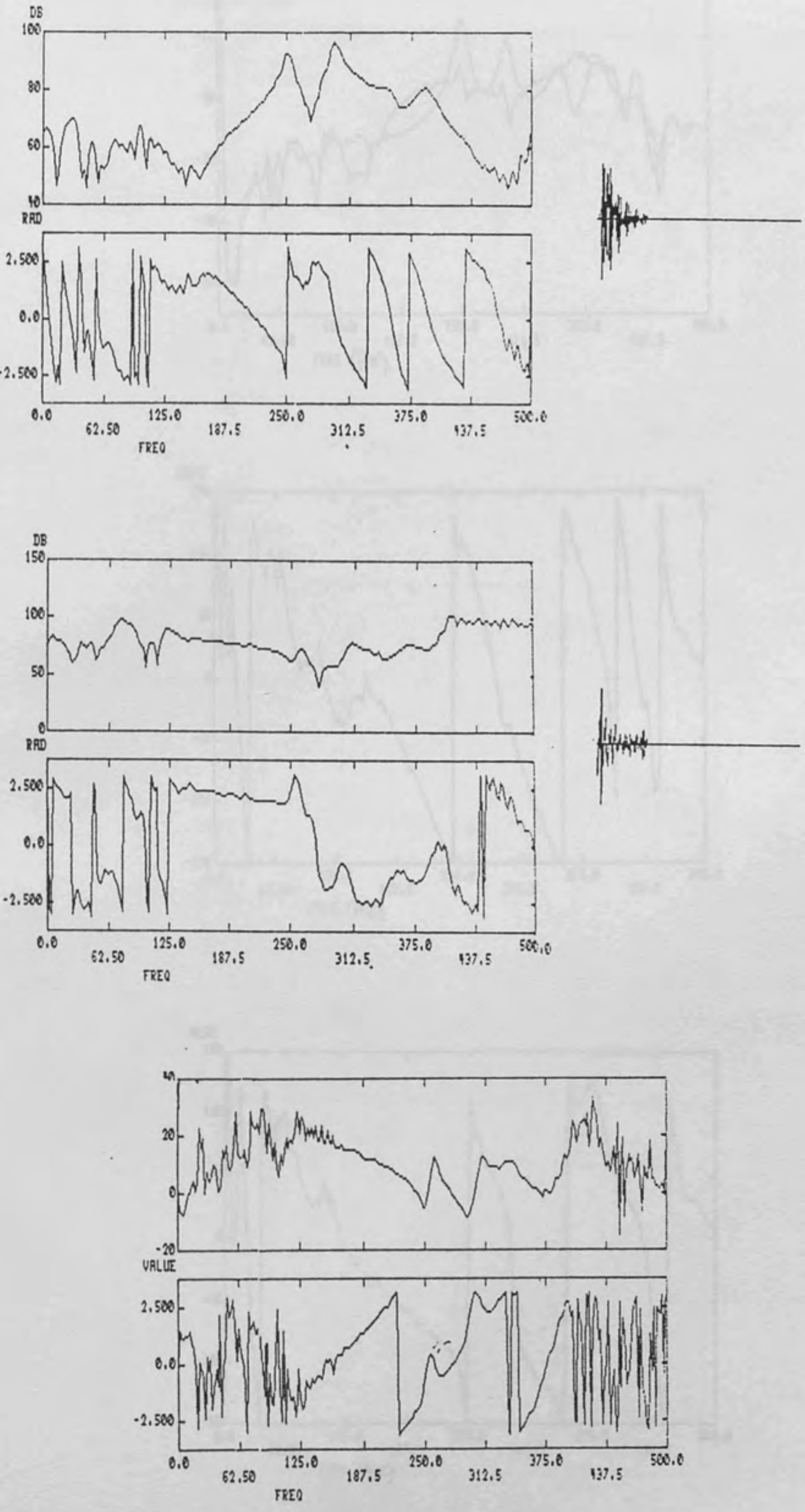
THEORETICAL ATTENUATION PLOTS AT 1,5 AND 10 KHZ

FIGURE 4.4.4 (ARBITRARY AMPLITUDE)



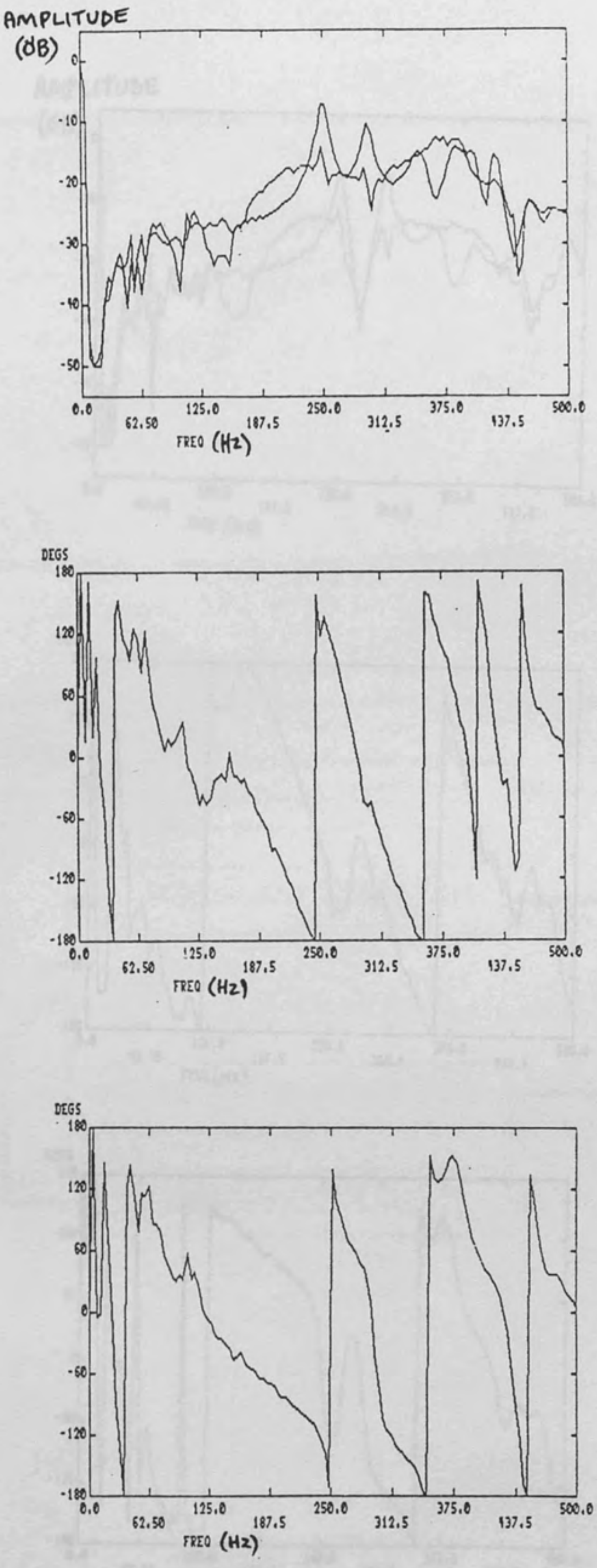
MEASUREMENTS ON THE SINGLE CHANNEL CONTROL SYSTEM
WITH THE MONITOR AT A DIFFERENT POSITION Y10,Y30,Y32 AND Y12

FIGURE 4.4.5 (ARBITRARY AMPLITUDE)



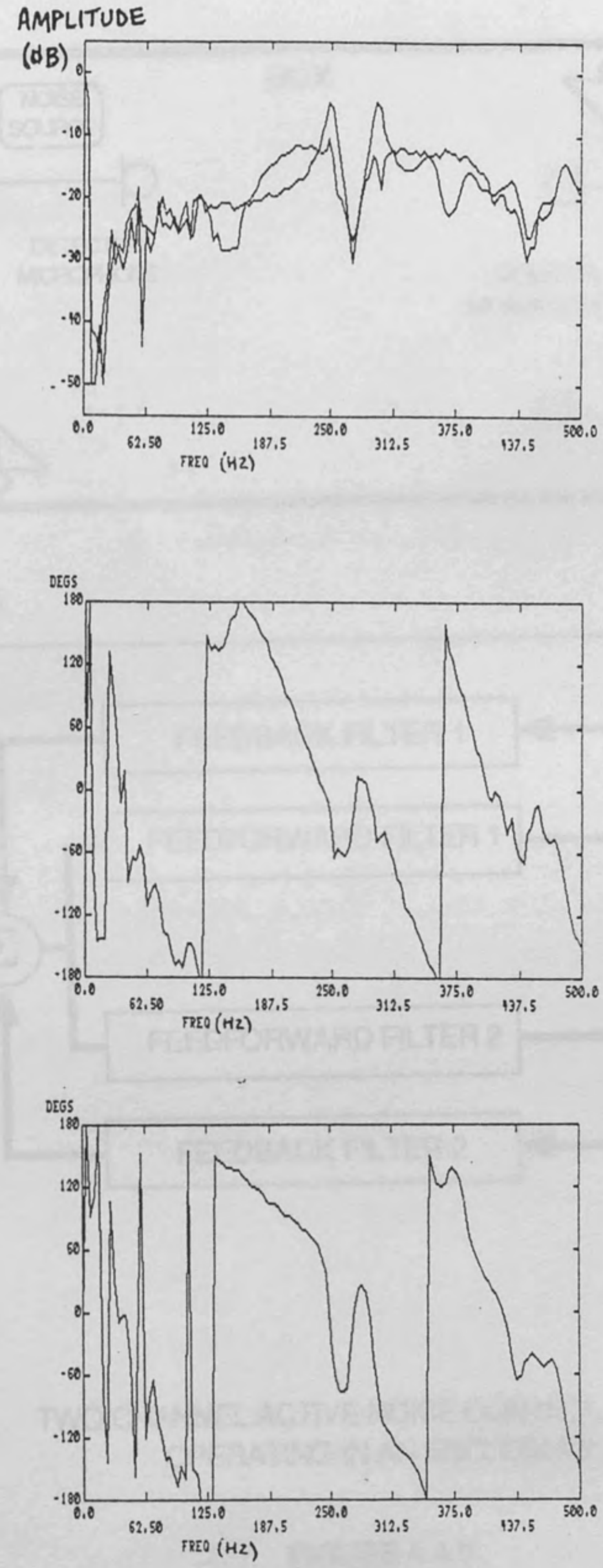
DERIVED FEEDBACK AND FEEDFORWARD FILTERS 128 POINT FIR
AND THE NON-CAUSAL SPECTRUM Y30/Y32 (ARBITRARY AMPLITUDE)

FIGURE 4.4.6

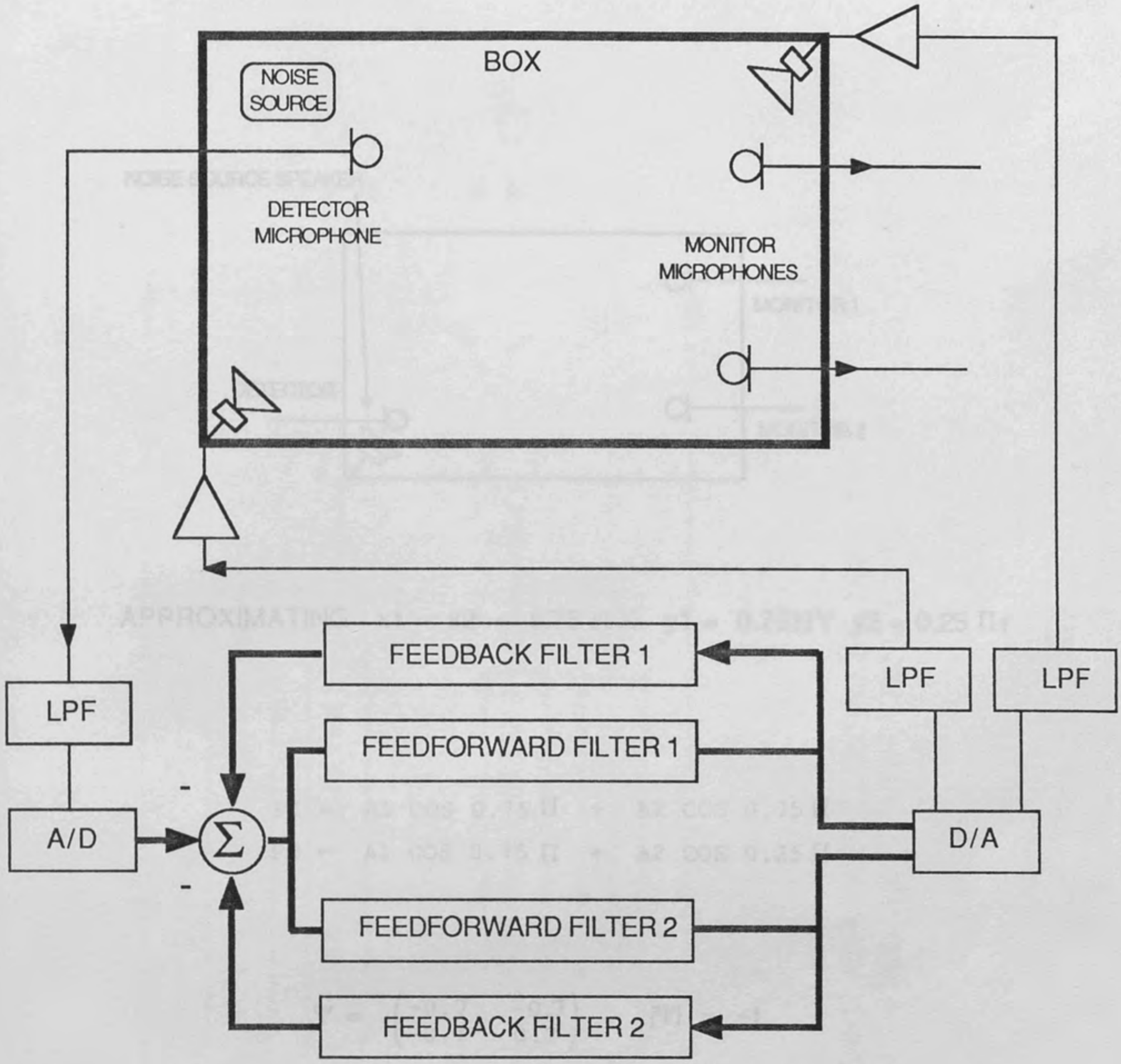


AMPLITUDE AND PHASE SPECTRA AT THE MONITOR MICROPHONE WITH AND WITHOUT THE CONTROL SYSTEM OPERATING

FIGURE 4.4.7

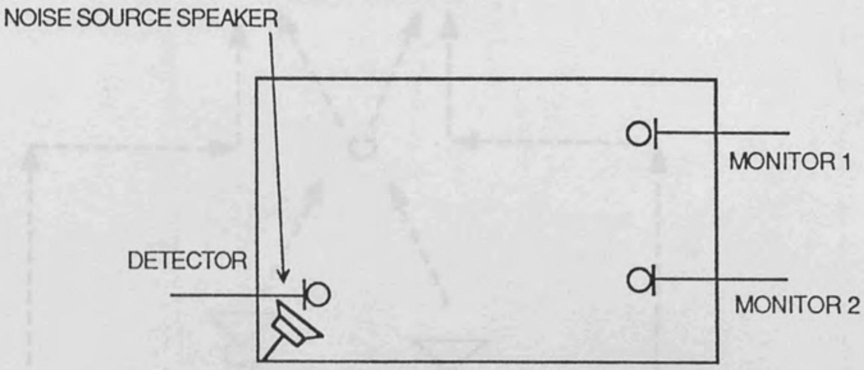


AMPLITUDE AND PHASE SPECTRA AT THE OBSERVER
MICROPHONE WITH AND WITHOUT THE CONTROL SYSTEM OPERATING
FIGURE 4.4.8



TWO CHANNEL ACTIVE NOISE CONTROL SYSTEM
OPERATING IN AN ENCLOSURE

FIGURE 4.4.9



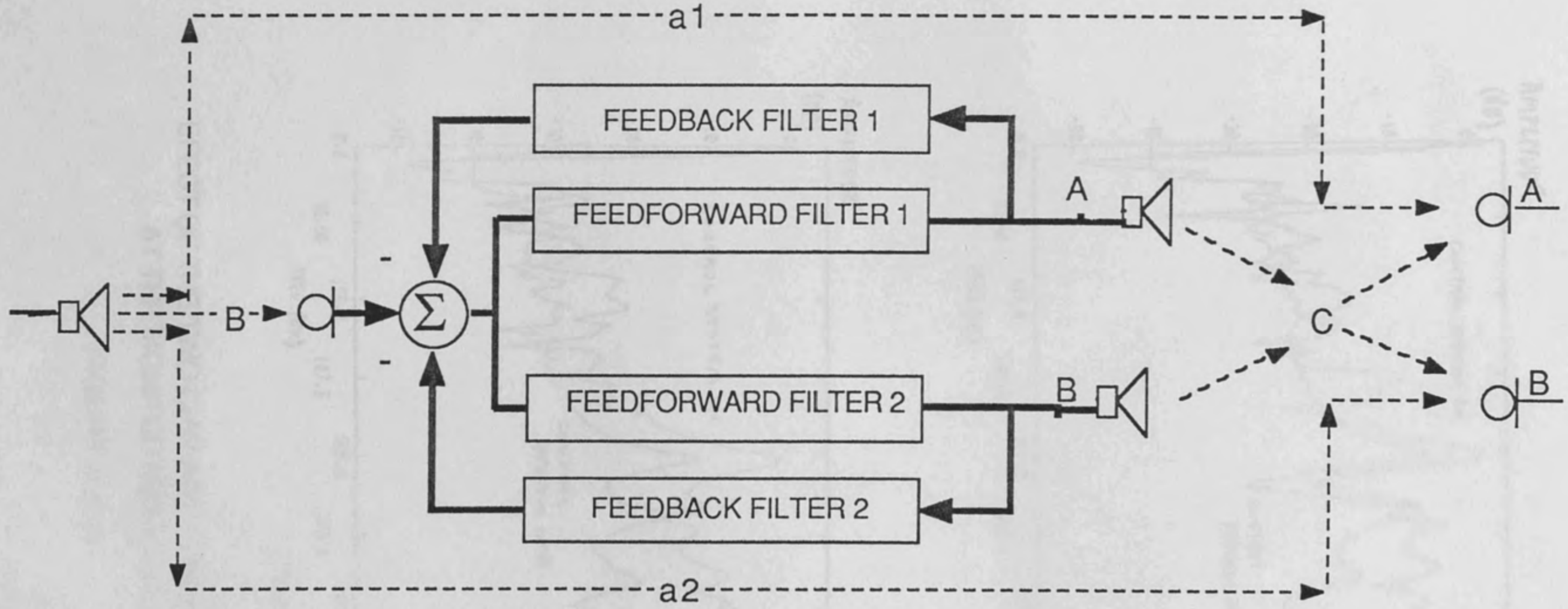
APPROXIMATING $x1 = x2 = 0.75\,\Pi X$ $y1 = 0.75\Pi Y$ $y2 = 0.25\,\Pi Y$

$$P1 = A1 \cos 0.75\,\Pi + A2 \cos 0.75\,\Pi$$
$$P2 = A1 \cos 0.75\,\Pi + A2 \cos 0.25\,\Pi$$

$$\Psi = \begin{pmatrix} -0.7 & -0.7 \\ -0.7 & 0.7 \end{pmatrix} \qquad |\Psi| = -1$$

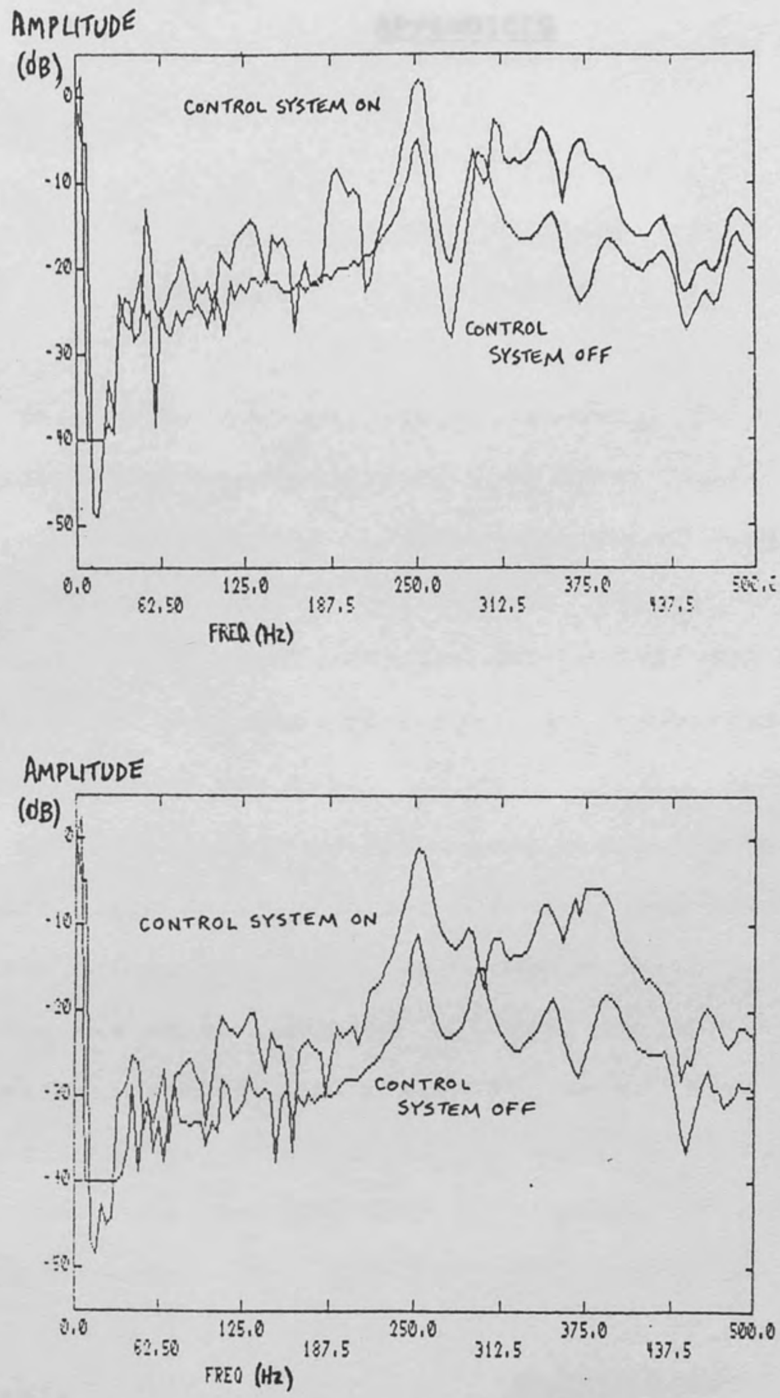
TWO CHANNEL ACTIVE CONTROL SYSTEM
SIDE PROJECTION

FIGURE 4.4.10



TWO CHANNEL ACTIVE NOISE CONTROL SYSTEM
SHOWING THE RELEVANT TRANSFER FUNCTIONS

FIGURE 4.4.11



EFECT OF THE TWO CHANNEL CONTROL SYSTEM
AT THE MONITOR MICROPHONES

FIGURE 4.4.12

APPENDICES

APPENDIX 1

MULTIPLEXER-DEMULTIPLEXER UNIT

INTERFACED TO THE TMS32020 BOARD.

This appendix describes the design and operation of the multiplexer-demultiplexer unit which was constructed to interface to the Loughborough Sound Images TMS32020 board. The unit has two input ports, both connected to the analogue to digital controller of the TMS32020 board. The digital to analogue converter is connected to the two output ports of the unit.

Control logic from the microprocessor determines that on each clock cycle each input channel is sampled and a sample is output from each output channel. The unit is constructed to be operated from the specific set of software instructions presented here. Even though two channels may not be required the unit still operates in this way and the data from the unwanted channel needs to remain unprocessed. Whatever processing needs to be performed is inserted into the relevant sections of the assembly language program.

The following technical specifications are presented;

A listing of the Texas TMS32020 assembly language program to run on the microprocessor and operate the two channel input and output unit.

A block diagram of the MDMU circuitry.

A timing diagram of the control logic signals.

A circuit diagram of the MDMU.

The operation of the unit may be understood by refering to the above while reading the description of the software below;

OUT TEMP,TIM The on-board interval timer is used; to use this sample clock source , link LK6a of the TMS32020 board must be inserted, and link LK6b must be absent (Ref. LSI).

RXF XF = 0 so mux switch looks at chana first

ISR EQU $ clock cycle initiates INTO signal which causes the interrupt service routine to start. clock sends SH high causing the two input sample and hold devices to hold; therefore the two input signals are captured at the instant of the clock pulse. The clock also initiates a conversion of the ADC (chana signal) such that chana is held on the ADC latches

SST1 STATUS load status register values into data memory address STATUS

LACK BIT4 load value of data memory address BIT4 (= >10) into accumulator

AND STATUS AND STATUS with accumulator to determine the value of the forth bit of the status register (which contains the value of XF)

BNZ NEXT branch to statement NEXT if accumulator (now containing the value of XF) is not equal to zero; initially XF = 0 so program continues to the next statement on the first pass through the program

SXF XF = 1 causes mux switch to link channel 2 through to ADC, next pass through the program causes it to jump to the NEXT instruction

CALL DELAY

CALL DELAY 10 microseconds delay to wait for switch moved by XF to settle before releasing the signal level from the sample and hold; need to wait sufficient time to be sure of seeing the desired channel at the switch and not part of the voltage level from the other channel

IN CHANA,3 causes the signal (chana) on the ADC gates to pass to the data bus; this is the signal whose conversion was initiated on the clock pulse. Because the ADC is being read as port 3 another conversion (chanb signal passing through the MUX switch) is initiated.

```
     EINT                 enable interrupt;

     RET                  return to HERE to wait for
interrupt; end of conversion signal from the ADC causes
INTO to go low which causes interrupt

 ....BNZ   NEXT           on this pass through the program
XF = 1 so program jumps to NEXT

NEXT IN    CHANB,2        causes the signal (chanb) on the
ADC gates to pass to the data bus. Because the ADC is
being read as port 2 the ADC just reads chanb and another
conversion is not iniatated (and no EOC signal occurs so
an interrupt doesnt occur)

     OUT   CHANA,2        output chana sample from DAC to
buffer

     OUT   CHANA,3        output chana sample from buffer
through switch to sample and hold of channel A (SHA) which
is on sample because initially XF = 1

     CALL  DELAY          wait for conversion to finish

     RXF                  XF = 0 which puts SHA on hold;
ie. holds chana output value

     RPT   3
     NOP                    1 microsecond delay before
```

outputting chanb sample from DAC. note; unlike the situation on the input side the system does not need to wait 10 microseconds after changing XF because in case the switch is switching in the direction of the signal flow; as long as the voltage level is present at the output from the DAC buffer it will be seen by the SHA device after the switch has settled

```
        OUT   CHANB,2        send chanb sample to DAC from bus
        OUT   CHANB,3        send chanb sample from DAC to
```

buffer; chana is output from the buffer

2 microseconds after XF is set to 0 then switch sends output from DAC buffer to SHB; this delay is present to ensure that the sample and hold device SHA is on hold before chanb is passed through the signal

SHB is put on hold by the pulse XFP going high

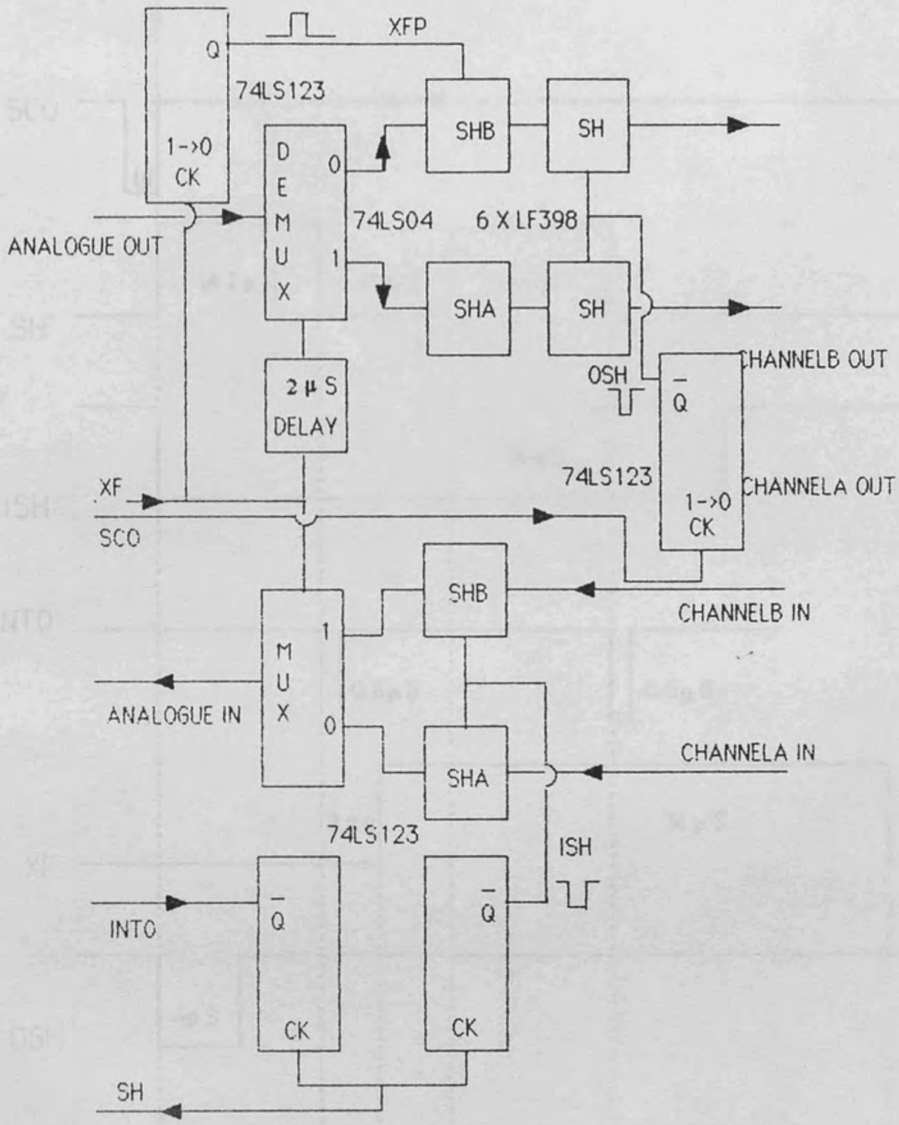
the samples at SHA and SHB have been held at different times so another set of sample and holds are needed to output both values at the same time

```
        EINT
        RET                  SCO going low from the clock
```

pulse causes the final set of sample and holds at the output to change to sample; in this state the output voltage levels are output to the output ports. after 4 microseconds the SH devices return to hold and those signal levels continue to be output until the next clock pulse.

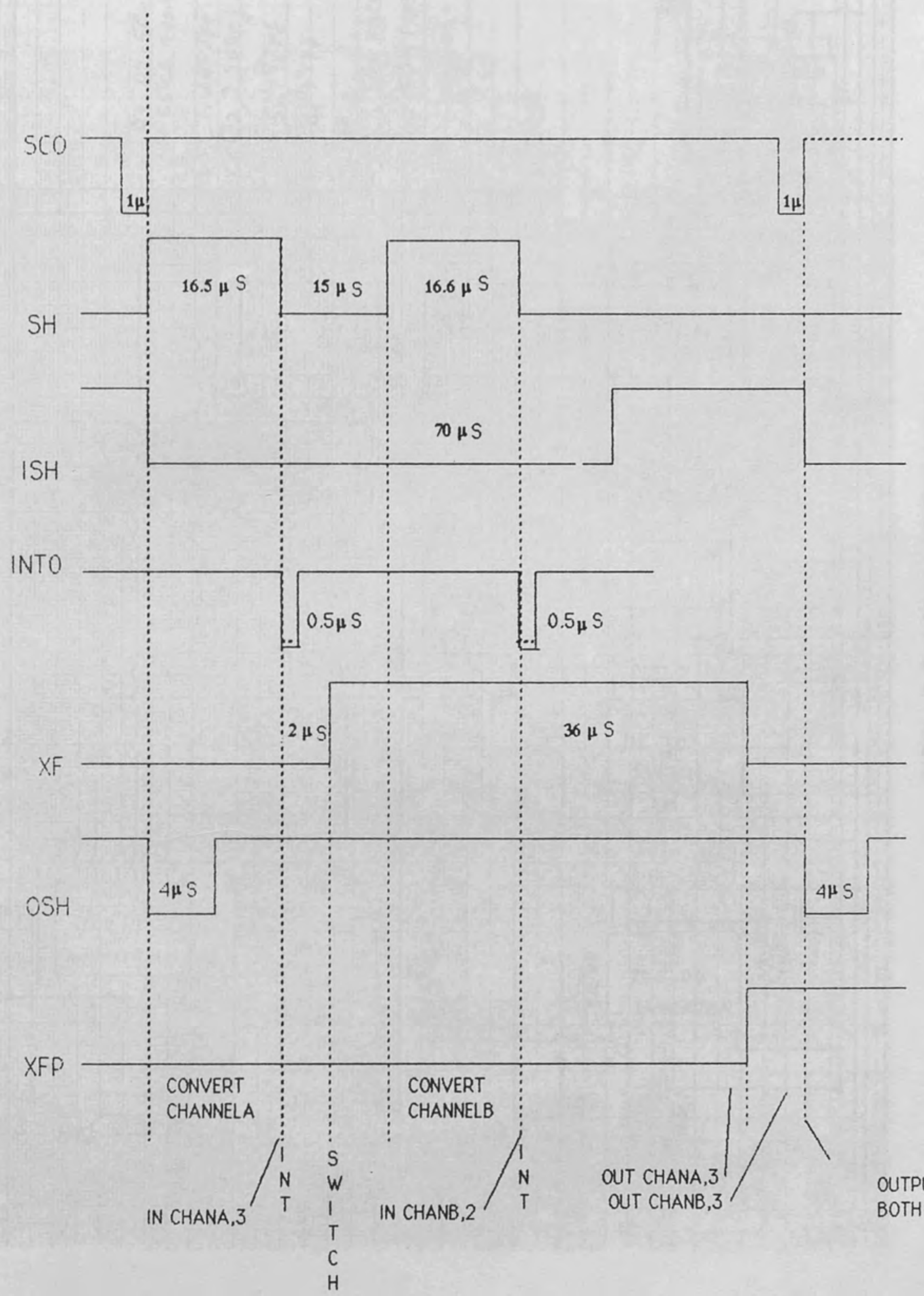
```
*  I/P & O/P TEST PROGRAM   40 dB INTERMODULATION
*   CHANA I/P TO CHANA O/P    BETWEEN CHANNELS
*   CHANB I/P TO CHANB O/P
PAGE0     EQU  0
IMASK     EQU  >FFC1         ENABLES INTERRUPT 0 & MASKS OFF OTHERS
IMR       EQU  4              INTERRUPT MASK REGISTER
TIM       EQU  1
BIT4      EQU  >10
TIMVAL    EQU  >EC7F         TO SET 1 KHz CLOCK RATE;
DUM       EQU  >63            TO SET A SAMPLING FREQUENCY OF N
TEMP      EQU  >64            TIMVAL = 1 - 5000000/N in HEX
STATUS    EQU  >65
CHANA     EQU  >66
CHANB     EQU  >67
************************
*
*
* INSERT EXTRA ADDRESS & DATA CONSTANTS IN HERE
*
*
************************
          AORG  0             BRANCH TO START ON RESET
          B     START
          AORG  2             BRANCH TO ISR ON INTO
          B     ISR
START     AORG  >400          START OF PROGRAM
*
************************
*
*
* DATA STORED IN PROGRAM MEMORY & INITIAL PROGRAM SET UP IN HERE
*
*
************************
* SET CLOCK RATE
          SAR   0,DUM
          PSHD  DUM
          LDPK  PAGE0
          LRLK  0,TIMVAL
          SAR   0,TEMP
          OUT   TEMP,TIM
* ENABLE INTERRUPT
          RXF                  XF = 0  SO SAMPLES CHANA FIRST
          LRLK  0,IMASK        STORE IMASK VALUE
          SAR   0,IMR          IN INTERRUPT MASK REGISTER
          POPD  DUM            TO ENABLE INTERRUPT 0
          LAR   0,DUM
          EINT                 CLOCK INITIATES CONVERSION FOR CHANA
HERE      B     HERE           EOC SENDS SH HIGH, ie. TO HOLD
*                              TRAILING EDGE OF EOC ENABLES INTO
************************
```

```
*************************
ISR      EQU   $              101 CYCLES = 20.2 MICROSECONDS
*                             MAX CLOCK RATE = 49500 Hz
*                             A READ IN ON XF = 0
*                             A READ IN ON XF = 1
*                             A PUT OUT ON XF = 1
*                             B PUT OUT ON XF = 0
         LDPK  PAGE0
         SST1  STATUS         CHECK STATUS OF XF
         LACK  BIT4
         AND   STATUS
         BNZ   NEXT           BRANCH TO NEXT IF XF = 1
         SXF                  SET XF = 1
         CALL  DELAY          WAIT FOR SWITCHES TO SETTLE
         CALL  DELAY
         IN    CHANA,3
*          READS IN CONVERSION INIATED ON CLOCK PULSE
*                             INITIATES CHANNEL 'B' CONVERSION
*          SECOND INTERUPT OCCURS WHEN CONVERSION FINISHED
*          SPENDS MOST TIME SENDING CHANA TO ADC
*                             30 MICROSECONDS SENDING CHANB
         EINT
         RET
NEXT     IN    CHANB,2      READS IN CHANNEL 'B' SAMPLE
*                           READS IN ONLY, NO INITIALISING
***************************
*                                    SAMPLING        PROCESSING TIME
*                                    FREQ (KHz)    AVAILABLE (MICROSEC
* ENTER MAIN PROGRAM ROUTINE HERE ------------------------------------
*                                       10                  79
*                                        5                 179
*                                        4                 229
*                                        2                 479
*                                        1                 979
***************************
         LDPK  PAGE0
         OUT   CHANB,2
         OUT   CHANB,3      OUTPUT CHANNEL 'A' SAMPLE
         CALL  DELAY        WAIT FOR EOC (DAC)
         RXF                RESET XF = 0
         RPTK  3            1 MICROSEC DELAY SO THAT SHA IS SURE TO
         NOP                HOLD CHANA OUTPUT FROM DAC
         OUT   CHANA,2
         OUT   CHANA,3      OUTPUT CHANNEL 'B' SAMPLE
*                           IMMEDIATELY AFTER XF SET TO 0
*                           DUE TO 4 MICROSEC PULSE ON XFP
*                           WHEN CHANB IS ADDRESSED
         EINT
         RET
*************************
DELAY    SAR   0,DUM
         PSHD  DUM
         LARK  0,4          25 CYCLES
INNER    LARP  0            5 MICRO SECOND DELAY
         BANZ  INNER
         POPD  DUM
         LAR   0,DUM
         RET
         END
```



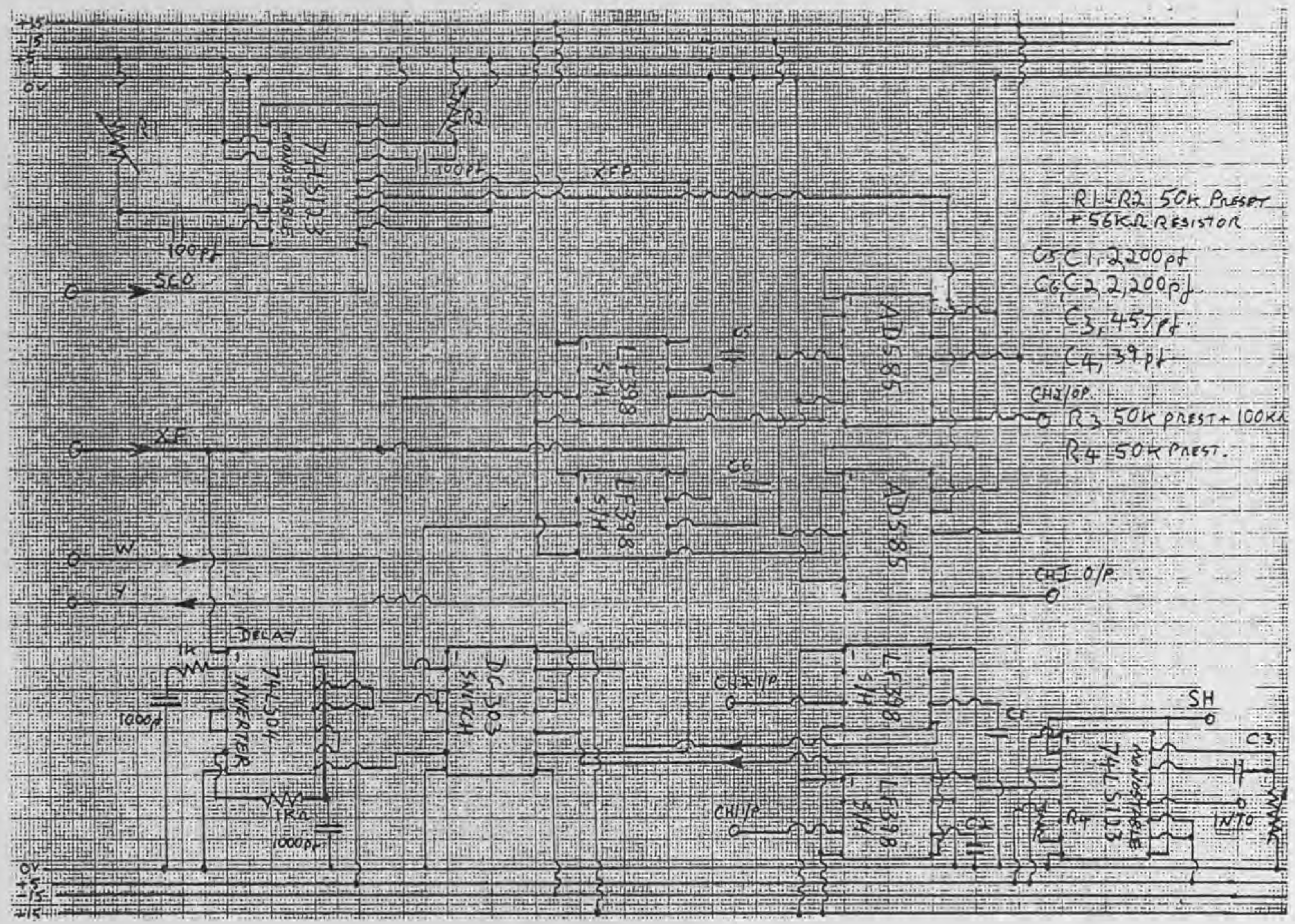
BLOGK DIAGRAM OF THE CIRCUITRY FOR THE MDMU

FIGURE A1.1



TIMING DIAGRAM FOR THE MDMU CONTROL LOGIC

FIGURE A1.2



CIRCUIT DIAGRAM OF THE MDMU

FIGURE A1.3

APPENDIX 2

Texas TMS32020 assembly language program to realise the single channel digital controller.

```
* SN RATIO DETERMINED BY 40 dB INTERMODULATION BETWEEN CHANNELS
*        IDT     'MFIR2A'
***************
*              *
*    MFIR2A    *
*              *
***************
*
*
* TO RUN ON LOUGHBOROUGH 32020 BOARD VIA MULTIPLEXED INPUT & OUTPUT
*
* C.BEAN 12/86
*
*********************************************************************
*
*    2 FIR FILTERS      ---  /2 --------  FFF  -X8----  X2 ---
* LENGTH-128 COEFFICIENTS              |               |
*                                      |               |
* SAMPLING FREQUENCY = 1 KHZ         ----  FBF  --<--
*
*************************************************************
*
*   CHANA I/P TO CHANA O/P
*   CHANB I/P TO CHANB O/P
*
PAGE0     EQU    0
IMASK     EQU    >FFC1     To enable interrupt 0
IMR       EQU    4
TIM       EQU    1
BIT4      EQU    >10
TIMVAL    EQU    >EC7F     For 1 KHz clock rate
DUM       EQU    >63
TEMP      EQU    >64
STATUS    EQU    >65
CHANA     EQU    >66
CHANB     EQU    >67
PAGE6     EQU    6         Page 6 is in B1, starting at >300
PAGE7    EQU      7        Page 7 is in B1, starting at >380
YN       EQU      >0       START OF PAGE6    >300
XN       EQU      >0       START OF PAGE7    >380
FN       EQU      >68      IN PAGE0
OLDX     EQU      >3FF     END OF PAGE 7
OLDY     EQU      >37F     END OF PAGE 6
*
************************
*
          AORG   0              BRANCH TO START ON RESET
          B      START
*
          AORG   2              BRANCH TO ISR ON INTO
          B      ISR
* FILTER COEFFICIENTS INITIALLY IN PROGRAM MEMORY
*         WITH FEEDFORWARD COEFFICIENTS FIRST
CTAB1  AORG      >400
```

```
DATA    -540
DATA     531
DATA    -757
DATA     671
DATA    -605
DATA     617
DATA    -617
DATA     478
DATA    -826
DATA     667
DATA    -683
DATA     798
DATA    -474
DATA     129
DATA     218
DATA    -966
DATA     966
DATA    -986
DATA     982
DATA    -429
DATA     -76
DATA     433
DATA   -1314
DATA     679
DATA    -998
DATA     133
DATA     135
DATA    -704
DATA     568
DATA   -1031
DATA     282
DATA    -556
DATA    -380
DATA     352
DATA    -908
DATA     271
DATA    -536
DATA    -445
DATA      96
DATA    -663
DATA     314
DATA    -389
DATA     198
DATA    -148
DATA    -683
DATA     386
DATA    -937
DATA      63
DATA    -544
DATA     228
DATA      18
DATA    -478
DATA     536
DATA    -384
DATA      15
```

```
DATA    -376
DATA     400
DATA     560
DATA      40
DATA     671
DATA     527
DATA     450
DATA    -237
DATA     118
DATA     728
DATA      17
DATA     147
DATA     581
DATA     740
DATA    -162
DATA    -638
DATA     589
DATA     390
DATA    -373
DATA     244
DATA    1199
DATA     104
DATA    -630
DATA      46
DATA     413
DATA    -150
DATA    -291
DATA    1838
DATA    1924
DATA     724
DATA     957
DATA    1003
DATA     929
DATA    -552
DATA      62
DATA    1678
DATA     361
DATA     -74
DATA      68
DATA    1314
DATA     -90
DATA   -1482
DATA    1732
DATA     371
DATA    -597
DATA    -324
DATA    -378
DATA     309
DATA   -2973
DATA    -564
DATA    1211
DATA   -1351
DATA    1396
DATA     210
DATA    1027
```

```
DATA    377
DATA  -2756
DATA   1420
DATA  -2100
DATA    -98
DATA   -139
DATA   -790
DATA   4218
DATA  -5570
DATA   5529
DATA  -4873
DATA   2395
DATA  -2088
DATA   -740
DATA   3177
DATA  -6389
DATA    728
DATA  -3882
DATA     84
DATA    -39
DATA    -52
DATA    206
DATA    143
DATA   -186
DATA     59
DATA    343
DATA     -9
DATA   -230
DATA    241
DATA    315
DATA   -133
DATA    -66
DATA    363
DATA    151
DATA   -135
DATA     99
DATA    206
DATA      1
DATA    100
DATA     74
DATA    -44
DATA    237
DATA    231
DATA   -209
DATA     51
DATA    513
DATA     69
DATA   -359
DATA    308
DATA    589
DATA   -263
DATA   -275
DATA    618
DATA    270
DATA   -422
```

```
DATA    109
DATA    468
DATA    -68
DATA    -58
DATA    220
DATA     14
DATA    129
DATA    247
DATA   -250
DATA    -96
DATA    504
DATA     55
DATA   -697
DATA    127
DATA    874
DATA   -513
DATA   -805
DATA    956
DATA    461
DATA  -1002
DATA     11
DATA    841
DATA   -224
DATA   -559
DATA    294
DATA    217
DATA   -279
DATA     77
DATA   -102
DATA   -330
DATA    422
DATA    241
DATA   -933
DATA    -78
DATA   1388
DATA   -543
DATA  -1608
DATA   1444
DATA    985
DATA  -1883
DATA   -175
DATA   1569
DATA   -260
DATA  -1300
DATA    369
DATA   1090
DATA   -835
DATA   -428
DATA    815
DATA   -569
DATA     60
DATA    785
DATA  -1169
DATA   -418
DATA   1883
```

```
        DATA   -399
        DATA  -2594
        DATA   1729
        DATA   2194
        DATA  -3014
        DATA   -766
        DATA   3037
        DATA   -119
        DATA  -2440
        DATA    181
        DATA   2833
        DATA  -1392
        DATA  -2267
        DATA   2899
        DATA   -481
        DATA  -1592
        DATA   2404
        DATA  -1601
        DATA   -867
        DATA   3086
        DATA  -1034
        DATA  -2774
        DATA   2227
        DATA   2663
        DATA  -3603
        DATA  -1257
        DATA   3440
        DATA   -821
        DATA  -1713
        DATA    641
        DATA    517
        DATA     45
        DATA      4
        DATA     -4
        DATA     27
        DATA     31
*
START   EQU     $
*
        SPM     0               PM = 0 FOR NO SCALING IN APAC
        SSXM                    SXM = 1 SIGN EXTN IN ACC
        SOVM                    SET OVERFLOW MODE
*
* LOAD FILTER COEFFICIENTS TO DATA MEMORY
*
        LDPK    PAGE0
        LARP    AR0
        LRLK    AR0,>200        POINT TO BLOCK B0
        RPTK    >7F             2 X 128 COEFFICIENTS
        BLKP    CTAB1,*+        PM 400-47F TO DM 200-27F
        RPTK    >7F
        BLKP    >480,*+
        CNFP                    USE B0 AS PROGRAM AREA
*       - ONLY DO THIS AFTER TRANSFERING COEFF TO DATA MEMORY
* SET CLOCK RATE
```

```
        SAR     O,DUM
        PSHD    DUM
        LDPK    PAGEO
        LRLK    O,TIMVAL
        SAR     O,TEMP
        OUT     TEMP,TIM
* ENABLE INTERRUPT
        RXF                             RXF - 0  SO CHANA IS SAMPLED FIRST
        LRLK    O,IMASK
        SAR     O,IMR
        POPD    DUM
        LAR     O,DUM
        EINT
HERE    B       HERE
*
************************
ISR     EQU     $                       398 CYCLES = 79.6 MICROSECONDS
*                                       MAX CLOCK RATE = 12.5 KHz
*    INPUT                              A READ IN ON XF = 0
*   SEQUENCE                            B READ IN ON XF = 1
*   --------                            A PUT OUT ON XF = 1
*                                       B PUT OUT ON XF = 0
        PUSH
        LDPK    PAGEO
        SST1    STATUS
        LACK    BIT4
        AND     STATUS
        BNZ     NEXT                    BRANCH TO NEXT IF XF = 1
        SXF                             SET XF = 1
        CALL    DELAY
        CALL    DELAY
        LDPK     PAGE7
        IN      XN,3
*     READS IN CONVERSION INIATED ON CLOCK PULSE
*
*
*
        POP
        EINT
        RET
*
NEXT    IN      CHANB,2             READS IN CHANNEL 'B' SAMPLE
*                                   READS IN ONLY, NO INITIALISING
        POP
**************************
*
       LDPK     PAGE7
*
       LAC      XN,15                 XN + FN ; DIVIDE XN BY 2
       LDPK     PAGEO                  TO AVOID POSSIBLE OVERFLOW
       ADDH     FN                     IF BOTH LARGE
       LDPK     PAGE7
       SACH     XN
*
```

```
*
*
        LRLK    AR1,OLDX         LOAD OLDEST VALUE OF INPUT SIGNAL
        LARP    AR1               INTO REGISTER IN PREPARATION
        MPYK    0                 FOR MULTIPLICATION WITH FIRST
        ZAC                       COEFFICIENT (IN PM >FF00)
        RPTK    >7F               AT START OF CONVOLUTION
        MACD    >FF00,*-         FEEDFORWARD FILTER
        APAC
        LDPK    PAGE6
        SACH    YN,4            YN X 8 (X2 TO RETURN TO Q15 FORMAT)
*                                BECAUSE COEFF DIVIDED BY 8
        LAC     YN,1
        LDPK    PAGE0
        SACL    CHANA           CHANA X 2 FOR OUTPUT
        LDPK    PAGE6
*
        LRLK    1,OLDY          LOAD OLDEST VALUE OF OUTPUT
        LARP    AR1              SIGNAL INTO REGISTER IN
        MPYK    0                PREPARATION FOR CONVOLUTION
        ZAC
        RPTK    >7F
        MACD    >FF80,*-        FEEDBACK FILTER
        APAC
        LDPK    PAGE0
        SACH    FN,1            ; FN BACK TO Q15 FORMAT
*
*                               ; GAIN - 1 WHEN DMA 47F  - +4096
*                                           & DMA 4FF  = +32767
*                               ; +32767 = ONE IN Q15 FORMAT

************************************************************
*       OUTPUT SEQUENCE
*
         LDPK   PAGE0
         OUT    CHANB,2
         OUT    CHANB,3      OUTPUT CHANNEL 'B' SAMPLE
         CALL   DELAY
         RXF                 RESET XF = 0
         RPTK   3
         NOP
         OUT    CHANA,2
         OUT    CHANA,3        OUTPUT CHANNEL 'A' SAMPLE
*
         EINT
         RET
*************************
DELAY    SAR    0,DUM
         PSHD   DUM
         LARK   0,4          25 CYCLES
INNER    LARP   0            5 MICRO SECOND DELAY
         BANZ   INNER
         POPD   DUM
         LAR    0,DUM
         RET
```

APPENDIX 3

Texas TMS32020 assembly language program to realise the two channel digital controller.

```
*              40 dB INTERMODULATION BETWEEN CHANNELS
*        IDT      'MFIR4'
***************
*             *
*    MFIR4    *
*             *
***************
* PROGRAM TO IMPLEMENT AN ACTIVE CONTROL SYSTEM CONSISTING
*  OF A SINGLE DETECTOR AND TWO CONTROL SOURCES
* C.BEAN 4.87
***********************************************************
*                            ---/2------------FF1-----------X2-----
*    4 FIR FILTERS                 |    |                 |
* LENGTH-128 COEFFICIENTS          |    ------FB1-------
*                                  |-----------FF2-----------X2-----
* SAMPLING FREQUENCY = 1 KHZ            |                 |
*                                       ------FB2-------
***********************************************************
*
*   YNA    YNB     >300              +5901  +6029
*   OLDYA  OLDYB   >37F              +6028  +6156
*
PAGE0     EQU   0
IMASK     EQU   >FFC1
IMR       EQU   4
TIM       EQU   1
BIT4      EQU   >10
TIMVAL    EQU   >EC7F
DUM       EQU   >63
TEMP      EQU   >64
STATUS    EQU   >65
CHANA     EQU   >66
CHANB     EQU   >67
PAGE6     EQU   6       Page 6 is in B1, starting at >300
PAGE7    EQU     7      Page 7 is in B1, starting at >380
YNA      EQU     >0     START OF PAGE6   >300
YNB      EQU     >0     START OF PAGE6   >300
XN       EQU     >0     START OF PAGE7   >380
FNA      EQU     >68    IN PAGE0
FNB      EQU     >69    IN PAGE0
OLDX     EQU     >3FF   END OF PAGE7
OLDYA    EQU     >37F   END OF PAGE6
OLDYB    EQU     >37F   END OF PAGE6
PYNA     EQU     +5901  TEMP PMA STORE FOR O/P TIME SERIES OF FIRST FILT

PYNB     EQU     +6029  TEMP PMA STORE FOR O/P TIME SERIES OF SECOND FII

UNITY    EQU     1
ONE      EQU     >6A
PMYNA    EQU     >6B
PMYNB    EQU     >6C
************************
         AORG   0             BRANCH TO START ON RESET
         B      START
         AORG   2             BRANCH TO ISR ON INT0
         B      ISR
CTAB1   AORG     >400
```

LIST OF 4 X 128 = 256 FILTER COEFFICIENTS

```
        DATA   2522
        DATA  -1274
        DATA  -2293
        DATA   2001
        DATA   2152
        DATA  -3374
        DATA  -1175
        DATA   3096
        DATA   -812
        DATA  -1641
        DATA    612
        DATA    540
        DATA     90
        DATA     34
        DATA     -3
        DATA     39
        DATA     23
*
*
START   EQU     $
*
        SPM     0                 PM = 0 SO NO SCALING IN APAC
        SSXM                      SXM = 1 SO SIGN EXTN IN ACC
        SOVM                      SET OVERFLOW MODE
* INITIALISE CONSTANTS
          LDPK  PAGE0
          LRLK  AR1,UNITY
          SAR   1,ONE
          LRLK  AR1,PYNA
          SAR   1,PMYNA
          LRLK  AR1,PYNB
          SAR   1,PMYNB
* SET TIMING
          SAR   0,DUM
          PSHD  DUM
          LDPK  PAGE0
          LRLK  0,TIMVAL
          SAR   0,TEMP
          OUT   TEMP,TIM
* ENABLE INTERRUPT
          RXF                    XF = 0  LOOKS AT CHANA FIRST
          LRLK  0,IMASK
          SAR   0,IMR
          POPD  DUM
          LAR   0,DUM
          EINT
HERE    B       HERE
*
***********************
*                                CAN ONLY CONVOLUTE B0 WITH B1
*   -----------------            B0 CONTAINS XN,YNA & YNB  :  COEFFS IN B1
*        INPUT                   1354 CYCLES = 270.8 MICROSECONDS
*       SEQUENCE                 MAX CLOCK RATE = 3.6 KHz
*   -----------------            A READ IN ON XF = 0
*                                A READ IN ON XF = 1
```

```
*                                   A PUT OUT ON XF = 1
*                                   B PUT OUT ON XF = 1
ISR       EQU    $
          PUSH
          LDPK   PAGE0
          SST1   STATUS             CHECK STATUS OF XF
          LACK   BIT4
          AND    STATUS
          BNZ    NEXT               BRANCH TO NEXT IF XF = 1
          SXF                       SET XF = 1
          CALL   DELAY              WAIT FOR SWITCHES TO SETTLE
          CALL   DELAY
          LDPK   PAGE7
          IN     XN,3
*
*
*
*
          POP
          EINT
          RET
*
NEXT      IN     CHANB,2
*
          POP
*************************
*
         LDPK     PAGE7
*
         LAC      XN,15          ; XN + FN ; DIVIDE XN BY 2
         LDPK     PAGE0
         ADDH     FNA            ; TO AVOID OVERFLOW
         ADDH     FNB
         LDPK     PAGE7
         SACH     XN        ISR HAS 78 CYCLES UP TO HERE
*
* LOAD FILTER COEFFICIENTS
*
         CNFD
         LDPK     PAGE0
         LARP     AR0
         LRLK     AR0,>200       ; POINT TO BLOCK B0
         RPTK     >7F            ; 2 X 128 COEFFICIENTS
         BLKP     CTAB1,*+       ; PM 400-47F TO DM 200-27F
         RPTK     >7F
         BLKP     >480,*+
         LAC      PMYNB            STORE YNB IN PM +6029 TO +6156
         LRLK     0,>300           RETRIEVE YNA FROM PM +5901 TO +6028
         RPTK     >7F
         TBLW     *+
         LAC      PMYNA
         LRLK     0,>300
         RPTK     >7F
         TBLR     *+
         CNFP                    ; USE B0 AS PROGRAM AREA
```

```
*         - ONLY USE CNFP AFTER TRANSFERING BCOEFF TO DATA MEMORY
*
        LDPK    PAGE7
        LRLK    AR1,OLDX
        LARP    AR1
        MPYK    0
        ZAC
        RPTK    >7F
        MACD    >FF00,*-
        APAC
        LDPK    PAGE6
        SACH    YNA,4          YN X 8 (X2 TO RETURN TO Q15 FORMAT)
*                              BECAUSE COEFF DIVIDED BY 8
        LAC     YNA,1
        LDPK    PAGE0
        SACL    CHANA         CHANA X 2 FOR OUTPUT
        LDPK    PAGE6
*
        LRLK    1,OLDYA
        LARP    AR1
        MPYK    0
        ZAC
        RPTK    >7F
        MACD    >FF80,*-
        APAC
        LDPK    PAGE0
        SACH    FNA,1         ; FN BACK TO Q15 FORMAT
*
*  549 CYCLES SINCE CNFD     ; GAIN = 1 WHEN DMA 47F  = +4096
*                                       & DMA 4FF  = +32767
*                            ; +32767 = ONE IN Q15 FORMAT
*
        LRLK    AR0,>380    134 CYCLES FOR THESE 4 LINES
        LARP    AR0
        RPTK    >7F
        BLKD    >381,*+
*
*
        CNFD
        LDPK    PAGE0
        LARP    AR0
        LRLK    AR0,>200
        RPTK    >7F
        BLKP    >500,*+
        RPTK    >7F
        BLKP    >580,*+
        LAC     PMYNA          STORE YNA IN PM +5901 TO +6028
        LRLK    0,>300         RETRIEVE YNB FROM PM +6029 TO +6156
        RPTK    >7F
        TBLW    *+
        LAC     PMYNB
        LRLK    0,>300
        RPTK    >7F
        TBLR    *+
        CNFP
```

```
*
        LDPK     PAGE7
        LRLK     AR1,OLDX
        LARP     AR1
        MPYK     0
        ZAC
        RPTK     >7F
        MACD     >FF00,*-
        APAC
        LDPK     PAGE6
        SACH     YNB,4              YN X 8 (X2 TO RETURN TO Q15 FORMAT)
*                                   BECAUSE COEFF DIVIDED BY 8
        LAC      YNB,1
        LDPK     PAGE0
        SACL     CHANB             CHANA X 2 FOR OUTPUT
        LDPK     PAGE6
*
        LRLK     1,OLDYB
        LARP     AR1
        MPYK     0
        ZAC
        RPTK     >7F
        MACD     >FF80,*-
        APAC
        LDPK     PAGE0
        SACH     FNB,1             ; FN BACK TO Q15 FORMAT
*
************************************
*   OUTPUT SEQUENCE
************************************  REST OF ISR HAS 44 CYCLES
         LDPK  PAGE0
         OUT   CHANB,2
         OUT   CHANB,3
         CALL  DELAY
         RXF                  RESET XF = 0
         RPTK  3
         NOP
         OUT   CHANA,2
         OUT   CHANA,3
*
*
*
         EINT
         RET
*************************
DELAY    SAR   0,DUM
         PSHD  DUM
         LARK  0,4            25 CYCLES
INNER    LARP  0              5 MICRO SECOND DELAY
         BANZ  INNER
         POPD  DUM
         LAR   0,DUM
         RET
         END
*
```

APPENDIX 4

Fortran program to numerically generate the transient test signal.

```
C       PROGRAM TO PRODUCE SWEPT SINE SIGNAL
C       ISWEPT ==> ARRAY WHICH WILL HOLD SIGNAL
C       LEN    ==> TOTAL NUMBER OF POINTS INCLUDING ZEROS
C       LN     ==> NUMBER OF ZEROS AT END
C       F1,F2  ==> START & END FREQUENCIES OF SWEEP
C       T      ==> TIME LENGTH OF SIGNAL
C       CLF    ==> SIGNAL OUTPUT RATE
C       ISTART,IEND  ==> NUMBER OF WINDOWED POINTS
        CHARACTER*12 FNAME
        DIMENSION ISWEPT(4000),SWEPT(4000)
   9    FORMAT(F12.6)
  28    FORMAT(I6)
  59    FORMAT(A)
        PI=3.14159265
        WRITE(*,'(A\)')' FILENAME SIGNAL TO BE STORED IN          : '
        READ(*,59) FNAME
        OPEN(41,FILE=FNAME,STATUS='NEW')
        WRITE(*,'(A\)')' START & END FREQS                        : '
        READ(*,*)F1,F2
        WRITE(*,'(A\)')' SIGNAL OUTPUT RATE (Hz)                  : '
        READ(*,*)CLF
        WRITE(*,'(A\)')' TOTAL NUMBER OF POINTS INCLUDING ZEROS : '
        READ(*,28)LEN
        WRITE(*,'(A\)')' NUMBER OF END ZEROS                      : '
        READ(*,28)LN
        LN=LEN-LN
        T=LEN/CLF
        WRITE(*,8)T
   8    FORMAT(1X,'LENGTH OF SIGNAL IS ',G10.5,'S')
        TT=T*(FLOAT(LN)/FLOAT(LEN))
        A=PI*(F2-F1)/TT
        B=PI*2*F1
        DO 50 I=1,LN
        TFS=(FLOAT(I)/FLOAT(LN))*TT
        SWEPT(I)=SIN(A*(TFS**2)+B*TFS)*32767
  50    CONTINUE
        WRITE(*,'(A\)')' NUMBER OF WINDOW POINTS AT START & END : '
        READ(*,*)ISTART,IEND
        IF((ISTART.EQ.0).AND.(IEND.EQ.0))GOTO 60
        CALL WINDOW(SWEPT,LN,ISTART,IEND)
  60    CONTINUE
        DO 70 I-1,LN
        ISWEPT(I)=IFIX(SWEPT(I))
  70    CONTINUE
        LNN-LN+1
        DO 80 I=LNN,LEN
        ISWEPT(I)=0
  80    CONTINUE
        WRITE(41,28) LEN
        WRITE(41,28) (ISWEPT(I),I=1,LEN)
        WRITE(41,28) LN
        END
C
        SUBROUTINE WINDOW(SWEPT,ILEN,ISTART,IEND)
C       PROGRAM TO PRODUCE A PLUS OR MINUS ONE COSINE WINDOW
```

```
C        OVER THE FIRST ISTART & LAST IEND POINTS OF AN ILEN POINT ARRAY
         DIMENSION SWEPT(4000)
         PI=3.14159265
         LAST=ILEN-IEND
         DO 200 I=1,ISTART
         FRACT=(FLOAT(I)/FLOAT(ISTART))*PI
         SWEPT(I)=((1.0-COS(FRACT))/2.0)*SWEPT(I)
  200    CONTINUE
         DO 210 I=1,IEND
         FRACT=(FLOAT(I)/FLOAT(IEND))*PI
         II=LAST+I
         SWEPT(II)=((1.0+COS(FRACT))/2.0)*SWEPT(II)
  210    CONTINUE
         RETURN
         END
```

APPENDIX 5

Texas TMS32020 assembly language program to measure the acoustic responses of the control system.

```
          IDT  'MINOAA'
*         PROG TO I/P 1900 WORDS OF DATA --
*         & O/P 1900 WORDS OF STORED DATA
*           -- MAX I/P = +/-10V --
*
*       ----- /16 ---->--I/P        O/P--------->
*
*         ________________________________________
*           TO RUN ;-
*                               FPM +2612 +5900 0
*                               SB 470
*           TO SAVE DATA ;-
*                               SPM +4000 +4512
*
*         ________________________________________
PAGE0     EQU  0
IMR       EQU  >4
IMASK0    EQU  >FFC1
CLCK      EQU  >EC79
BIT4      EQU  >10
*
* DATA MEMORY LOCATIONS
*
DUM       EQU  >63
ONE       EQU  >64
ZERO      EQU  >65
NUMAVG    EQU  >66   NUMBER OF TIMES MEASUREMENT REPEATED
SIGOUT    EQU  >67   START OF OUTPUT SIGNAL
SIGIN     EQU  >68   START OF INPUT SIGNAL
STATUS    EQU  >69
DPOUT     EQU  >6A   OUTPUT DATA POINT
DPIN      EQU  >6B   INPUT DATA POINT
TEMP      EQU  >6C
CHANA     EQU  >6D
CHANB     EQU  >6E
*
          AORG  0
          B     ENTRY
          AORG  2
          B     ISR
*
TBLE      AORG  >400
          DATA  +1       ONE
          DATA  +0       ZERO
          DATA  +16      NUMBER OF MEASUREMENTS
          DATA  +2100    PMA OF START OF O/P SIGNAL
          DATA  +4000    PMA OF START OF I/P SIGNAL
*                        PMA = PROGRAM MEMEORY ADDRESS
* TRANSFER DATA VALUES TO RELEVANT DATA MEMORY ADDRESSES
*
ENTRY     EQU   $
          LARP  0
          LRLK  0,ONE
          RPTK  >4
          BLKP  TBLE,*+
```

```
             CNFP             CNF = 1 BLOCK B0 = PROG MEM
             SSXM             SXM = 1 SIGN EXTN IN ACC
             SPM    0         PM = 0 NO SHIFTS IN ADDITION
* SET CLOCK RATE
             LDPK   PAGE0
             LRLK   0,CLCK
             SAR    0,TEMP
             OUT    TEMP,1
* SET COUNTING CONSTANTS
             LAC    SIGOUT
             LARK   0,18
             LARK   1,99
* ENABLE INTERRUPT
             RXF              XF = 0   FIRST SAMPLE IS FROM CHANNEL A
             LRLK   2,IMASK0
             SAR    2,IMR
             EINT
LOOP         B      LOOP
*
ISR          EQU    $
* INPUT SEQUENCE
             PUSH
             SST1   STATUS
             LACK   BIT4
             AND    STATUS
             BNZ    NEXT
             SXF
             CALL   DELAY
             CALL   DELAY
             CALL   DELAY
             IN     DPIN,3
*
             POP
             EINT
             RET
*
NEXT         IN     CHANB,2
*
*************************************
*
             LAC    NUMAVG         CHECK IF MEASUREMENTS COMPLETED
             BZ     DISP
             POP
             TBLR   DPOUT          FETCH DATA POINT TO BE OUTPUT
             PUSH
*
             LAC    DPIN,12        DIVIDE I/P POINT BY 16 TO
             SACH   DPIN,0          ACHIEVE AVERAGING
             POP
             LT     ONE
             MPYK   1900
             APAC
             TBLR   TEMP           FETCH CURRENT SUM OF THIS
             PUSH                   DATA POINT
             LAC    TEMP
```

```
          ADD   DPIN          ADD CURRENT I/P POINT TO
          SACL  DPIN           CURRENT SUM
          POP
          TBLW  DPIN          WRITE CURRENT SUM TO PROG MEM
          ADD   ONE
          LT    ONE
          MPYK  1900
          SPAC
          LARP  1
          BANZ  JUMP        CHECK IF 100 CYCLES OF AR1 DONE
*
          LARK  1,99
          LARP  0
          BANZ  JUMP        CHECK IF 19 CYCLES OF AR2 DONE
*
          LARK  1,99
          LARK  0,18
          LAC   NUMAVG
          SUB   ONE
          SACL  NUMAVG
          BZ    FIN       CHECK IF MEASUREMENT DONE 16 TIMES
          LAC   SIGOUT
*
* OUTPUT SEQUENCE
*           NEED TO OUTPUT A ZERO AT END OF SEQUENCE
*            TO PREVENT DC OFFSET BEING O/P AT END

JUMP      OUT   CHANA,2
          OUT   CHANA,3
          CALL  DELAY
          RXF
          RPTK  3
          NOP
          OUT   DPOUT,2
          OUT   DPOUT,3
          EINT
          RET
*
DELAY     SAR   0,DUM
          PSHD  DUM
          LARK  0,4
INNER     LARP  0            5 MICROSECOND DELAY
          BANZ  INNER
          POPD  DUM
          LAR   0,DUM
          RET
*
FIN       NOP
          NOP
          NOP
          NOP
          NOP
          NOP
          NOP
```

```
          NOP
          NOP
          NOP
          NOP
          NOP
          NOP
*
* TO OUTPUT CAPTURED SIGNAL
*
          LAC   SIGIN
          LARK  0,18
          LARK  1,99
          PUSH
*
DISP      RXF
          RPTK  3
          NOP
          POP
          TBLR  DPOUT
          OUT   DPOUT,3
          ADD   ONE
          LARP  1
          BANZ  BOUNCE
          LARK  1,99
          LARP  0
          BANZ  BOUNCE
          LARK  1,99
          LARK  0,18
          LAC   SIGIN
BOUNCE    EINT
          RET
          END
```

APPENDIX 6

Texas TMS32020 assembly language program to measure the open loop response of the single channel control system.

```
        IDT         'MOLTFA'
***************
*           *
*   MOLTFA  *
*           *
***************

* PROGRAM TO INPUT A SIGNAL TO THE OPEN LOOP OF THE
* CONTROL SYSTEM AND CAPTURE THE RESPONSE
* C.BEAN 3/87
*
*************************************************************
*
* 2 FIR FILTERS           ----------->-O/P  I/P->-/16----FFF-X8>----------
* LENGTH-128 COEFFICIENTS            ¦                                   ¦
* SAMPLING FREQUENCY = 1 KHZ     ----------  FBF --<----------
*
************************************************************
*
*                                   40 dB INTERMODULATION
*   CHANA I/P TO CHANA O/P
*   CHANB I/P TO CHANB O/P
PAGE0   EQU   0
IMASK   EQU   >FFC1
IMR     EQU   4
TIM     EQU   1
BIT4    EQU   >10
TIMVAL  EQU   >EC7F
DUM     EQU   >63
TEMP    EQU   >64
STATUS  EQU   >65
CHANA   EQU   >66
CHANB   EQU   >67
SIG     EQU    >69      DATA LOCATION OFSTART OF INPUT SIGNAL
ONE     EQU    >6A
PAGE6   EQU    6        Page 6 is in B0, starting at >300
PAGE7   EQU    7        Page 7 is in B1, starting at >380
YN      EQU    >0       START OF PAGE6   >300
XN      EQU    >0       START OF PAGE7   >380
FN      EQU    >68      IN PAGE0
OLDX    EQU    >3FF     END OF PAGE7
OLDY    EQU    >37F     END OF PAGE6
UNITY   EQU    1
SIGIN   EQU    +4000    PROGRAM MEMORY ADDRESS OF START OF I/P SIG
*
************************
*
        AORG  0              BRANCH TO START ON RESET
        B     START
        AORG  2              BRANCH TO ISR ON INTO
        B     ISR
*
************************
CTAB1   AORG    >400
```

LIST OF 2 X 128 = 256 FILTER COEFFICIENTS

```
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
        DATA     0
*
START   EQU     $
*
        SPM     0                  PM = 0 SO NO SCALING IN APAC
        SSXM                       SXM = 1 SO SIGN EXTN IN ACC
        SOVM                       SET OVERFLOW MODE
* LOAD COEFFICIENTS INTO DATA MEMEORY
        LDPK    PAGE0
        LARP    AR0
        LRLK    AR0,>200           POINT TO BLOCK B0
        RPTK    >7F                2 X 128 COEFFICIENTS
        BLKP    CTAB1,*+           PM 400-47F TO DM 200-27F
        RPTK    >7F
        BLKP    >480,*+
        CNFP                       CNF = 1 FOR MACD
*       - ONLY DO CNFP AFTER TRANSFERING COEFF TO DATA MEMORY
* SET UP CONSTANTS IN RELEVANT DMA
        LRLK    AR1,SIGIN
        SAR     1,SIG            ; START OF I/P SIG AT PMA +4000
```

```
        LRLK    AR1,UNITY
        SAR     1,ONE
* INITIAL CONDITIONS & PROGRAM SET UP
*
        LRLK    AR1,0
        SAR     1,FN            ; INITIALLY SET FEEDBACK VALUE TO 0
        LAC     SIG             ; TBLW PUTS FIRST CAPTURED SAMPLE IN
*                                 PMA SPECIFIED BY ACCUMULATOR
        LARK    AR0,15          ; O/P & I/P REPEATED 16 TIMES
        LARK    AR2,18          ; 19 X 100 = 1900 SAMPLES
        LARK    AR3,99
************************
* SET UP TIMING
         SAR    0,DUM
         PSHD   DUM
         LDPK   PAGE0
         LRLK   0,TIMVAL
         SAR    0,TEMP
         OUT    TEMP,TIM
* ENABLE INTERRUPT
         RXF                    XF = 0  LOOKS AT CHANA FIRST
         LRLK   0,IMASK
         SAR    0,IMR
         POPD   DUM
         LAR    0,DUM
         EINT
HERE     B      HERE
*
************************
ISR      EQU    $               400 CYCLES = 80+ MICROSECONDS
*         -------               A READ IN ON XF = 0
*          INPUT                A READ IN ON XF = 1
*         ROUTINE               A PUT OUT ON XF = 1
*        ---------              B PUT OUT ON XF = 0
         PUSH
         LDPK   PAGE0
         SST1   STATUS          CHECK STATUS OF XF
         LACK   BIT4
         AND    STATUS
         BNZ    NEXT            BRANCH TO NEXT IF XF = 1
         SXF                    SET XF = 1
         CALL   DELAY           WAIT FOR SWITCHES TO SETTLE
         CALL   DELAY
         LDPK   PAGE7
         IN     XN,3
*
*
*
*
*
         POP
         EINT
         RET
NEXT     IN     CHANB,2      READS IN CHANNEL 'B' SAMPLE
*                            READS IN ONLY, NO INITIALISING
```

```
        POP
**************************
        LDPK    PAGE7
*
        PUSH
        LAC     XN              XN + FN
        LDPK    PAGE0
        ADD     FN
        LDPK    PAGE7
        SACL    XN
        POP
*
* THE AVERAGING ; BY ADDING 16 CAPTURED SAMPLES
        LDPK    PAGE0
        TBLR    TEMP
        PUSH
        LAC     TEMP
        LDPK    PAGE7
        ADD     XN
        SACL    XN
        POP
*
        TBLW    XN              ; WRITE SUMMED CAPTURED VALUE
        LDPK    PAGE0
        LT      ONE
        LDPK    PAGE7
        MPYK    1900
        SPAC                    ; DECREASE ACC TO LOOK AT O/P SAMPLE
        TBLR    XN              ; READ SAMPLE TO BE O/P
        LDPK    PAGE0
        LT      ONE
        LDPK    PAGE7
        MPYK    1901
        APAC                    ; INCREASE ACC
*
        PUSH
        LAC     XN,12           ; DIVIDE XN BY 16
        SACH    XN                TO STORE AVERAGE OF 16 MEASUREMENTS
        LRLK    AR1,OLDX
        LARP    AR1
        MPYK    0
        ZAC
        RPTK    >7F
        MACD    >FF00,*-
        APAC
        LDPK    PAGE6
        SACH    YN,4            YN X 8 (& X2 TO RETURN TO Q15 FORMAT)
*                                 BECAUSE COEFF DIVIDED BY 8
        LAC     YN,0
        LDPK    PAGE0
        SACL    CHANA
        LDPK    PAGE6
*
        LRLK    AR1,OLDY
        LARP    AR1
```

```
        MPYK    0
        ZAC
        RPTK    >7F
        MACD    >FF80,*-
        APAC
        LDPK    PAGE0
        SACH    FN,1        ; FN BACK TO Q15 FORMAT
        POP
*
        LARP    AR3
        BANZ    JUMP
        LARK    AR3,99
        LARP    AR2
        BANZ    JUMP
        LARK    AR3,99
        LARK    AR2,18
        LAC     SIG
        LARP    AR0
        BANZ    JUMP
        B       FIN
***************************
JUMP     LDPK   PAGE0
         OUT    CHANB,2
         OUT    CHANB,3
         CALL   DELAY
         RXF
         RPTK   3
         NOP
         OUT    CHANA,2
         OUT    CHANA,3
*
*
         EINT
         RET
**************************
DELAY    SAR    0,DUM
         PSHD   DUM
         LARK   0,4          25 CYCLES
INNER    LARP   0            5 MICRO SECOND DELAY
         BANZ   INNER
         POPD   DUM
         LAR    0,DUM
         RET
FIN     NOP
        NOP
        NOP
        NOP
        NOP
        NOP
        NOP
        END
* BEFORE RUNNING ---
*
*    LPM MOLTFA OR TEMP3
*    LPM X2     OR SIG
*    FDM 300  3FF  0
*    FPM +2612  +5900 0
```

APPENDIX 7

C.Bean and S.J.Flockton. Active control of acoustic noise in a small enclosure. Paper presented by the author at the Institute of Acoustics annual meeting, Cambridge; April 1988.

ACTIVE CONTROL OF ACOUSTIC NOISE IN A SMALL ENCLOSURE

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(1) Now with B&W Loudspeakers, Worthing

INTRODUCTION

Any sound field can be thought of as the sum of propagating and reverberant fields. Much study has concerned the control of propagating fields; particularly in ducts. However, the control of the propagating field inside a room is much more complicated due to the multiple reflections involved and fundamentally requires that the control speakers be positioned in the line of the propagating field. Indeed, a propagating field is best controlled with a secondary source positioned as close as possible to the noise source. However, with multiple or large sources this may be difficult or impossible.

The principle of superposition can be applied to linear sound fields enabling the direct and reverberant fields to be considered separately. This paper is concerned with the active control of the low order modes of a reverberant field. In real situations the acoustic wave pattern of a reverberant field will be complicated, due to the shape of the enclosure and to objects and people within the enclosure, but a useful underastanding of the problem may be obatined by studying the simple stuation of a rectangular enclosure. A number of papers have been published concerning the active control of harmonic sound fields. Nelson [1] has shown that substantial reductions in the net acoustic power radiated can be achieved if the control sources are within half a wavelength of the noise source. Bullmore [2] has extended this theory to sound fields of low modal density by minimising the sum of the squared pressures at a number of different sensor locations and has shown that attenuation close to optimum can be achieved. It has also been shown how attenuation can be achieved with control sources separated from the noise source by distances of greater than half a wavelength. Little material has been published concerning experiments on the active control of broadband noise within an enclosure.

This paper describes the implementation of a system for the active control of the low order modes of the reverberant field in a small enclosure (where "small" infers that only a small number of acoustic modes dominate the field)

THEORY

In order to attenuate globally a sound field or produce a volume of attenuation it is necessary that the monitoring positions are chosen to be representative of the sound field throughout the volume of interest. In the case of a reverberant field it is necessary that the microphones pick up sufficient information about the dominant modes of the field. Let the sound field in an enclosure be dominated by n modes and the amplitude of the i'th mode be $A_i(t)$. Let the pressure in the enclosure be sensed by n sensors and the pressure at the j'th sensor be $P_j(t)$. Then the pressures at sensors 1 and 2 will be

$$P_1(t) = \Psi_{11} A_1(t) + \Psi_{21} A_2(t) + \ldots + \Psi_{n1} A_n(t)$$

$$P_2(t) = \Psi_{12}A_1(t) + \Psi_{22}A_2(t) + \ldots + \Psi_{n2}A_n(t)$$

where Ψ_{ij} is the characteristic function of the i'th mode at the j'th sensor position. It represents the fraction of the standing wave present at a position: $\Psi = 0$ at a node and is a maximum at an anti-node. The equations can be represented in matrix form:

$$\boldsymbol{P} = \Psi \boldsymbol{A}$$

and the modal pressures at a point obtained from the inverse equation

$$\boldsymbol{A} = \Psi^{-1}\boldsymbol{P}$$

Therefore in principle the characteristic functions (or eigenfunctions) of the modes need to be known in order to determine the modal pressures. Some knowledge of the mode shapes is also necessary when determining appropriate monitor positions. The important consideration in choosing the monitor positions is that the information present in the signals from the sensors is sufficient to define adequately all the modal amplitudes within the working range of the control system. Each monitor needs to be placed in an independent position from the others such that the simultaneous equations presented above can be solved.

An example will illustrate the meaning of the term independent. In practice it will be desirable to monitor a mode at or near an antinode to maximise the pressure detected. However, consider the case of monitoring the 1,0 and 0,1 modes in a 2-dimensional rectangular enclosure at positions in diagonally opposite corners. The matrix Ψ is then equal to $\left[\begin{smallmatrix}-1 & 1\\ 1 & -1\end{smallmatrix}\right]$, and as this matrix is singular, i.e. the determinant is zero, it cannot be inverted and hence the modal pressures cannot be resolved. This has occured because the chosen monitor positions were not independent; each position detected the same component of each mode. Note, however, that it is not necessary for the monitors to determine the modal amplitudes completely, only that there is sufficient information about the modal amplitudes present in the signals to avoid its being swamped by interfering noise.

MULTICHANNEL CONTROL SYSTEM

An active system consisting of a number of detectors and sources capable of controlling the field at a number of monitor positions is shown in figure 1. The letters in the figure are matrices of frequency responses between the elements. It has been shown [3] how the responses of the controllers needed between the detectors and sources are given by

$$\boldsymbol{T} = (\boldsymbol{C}^{\mathrm{H}}\boldsymbol{E}\,\boldsymbol{F} - \boldsymbol{C}^{\mathrm{H}}\boldsymbol{C})^{-1}\boldsymbol{C}^{\mathrm{H}}\boldsymbol{E}$$

where

- $\boldsymbol{T}$ is the matrix of the transfer functions of the controller needed to give optimum attenuation at the monitors,
- $\boldsymbol{C}$ is the matrix of the transfer functions between the control sources and the monitors,
- $\boldsymbol{F}$ is the matrix of the transfer functions of the acoustic feedback paths between the control sources and the detectors,
- $\boldsymbol{A}$ is the matrix of the transfer functions between the noise sources and the monitors,
- $\boldsymbol{B}$ is the matrix of the transfer functions between the noise sources and the detectors, and

E is the matrix of the transfer functions between the detectors and the monitors, and is equal to AB^{-1}.

Single Detector, Single Secondary Source, Two Monitors

Consider a controller consisting of one detector and one source controlling the field at two monitor positions . The required controller is given by

$$t = \left(\begin{bmatrix} c_1^* & c_2^* \end{bmatrix} \begin{bmatrix} e_1 \\ e_2 \end{bmatrix} f - \begin{bmatrix} c_1^* & c_2^* \end{bmatrix} \begin{bmatrix} c_1 \\ c_2 \end{bmatrix} \right)^{-1} \begin{bmatrix} c_1^* & c_2^* \end{bmatrix} \begin{bmatrix} e_1 \\ e_2 \end{bmatrix}$$

where the matrices T, C, E and F have components t, c_1, c_2, e_1,e_2 and f. Multiplying out the matrices leads to

$$t = \frac{c_1^* e_1 + c_2^* e_2}{(c_1^* e_1 + c_2^* e_2) f - (c_1^* c_1 + c_2^* c_2)}$$

and extending this to a controller with n monitors gives

$$t = \frac{1}{f - \dfrac{\sum_{i=1}^{n} c_i^* c_i}{\sum_{i=1}^{n} c_i^* e_i}}$$

Simple control theory indicates that this can be implemented with a pair of electronic filters; one between the detector and the source in parallel with another cancelling the acoustic feedback from the source to the detector. Increasing the number of or moving the monitor microphones does not affect the acoustic feedback in the system. Hence designing a controller in this way, with independent feedback compensation, means that the monitors can be moved without altering the feedback incorporated in the controller.

Two Detectors, Two Secondary Sources

Consider the general arrangement of figure 2. The acoustic feedback paths add together at the detector microphone (actually the point of entry to the digital system). This feedback can be counteracted by modelling each acoustic path electronically and summing the electronic feedback paths at an equivalent position to the acoustic feedback paths. The success of the method lies in the simple topology of the multichannel controller, the simplicity is rendered by the positions where the feedback paths meet, namely *before* the signal splits to enter the separate feedforward paths to the speakers. The advantage of this configuration is that the electronic feedback filter has a simple response which only needs to model the acoustic feedback due to that channel alone. These paths are also causal ensuring that they can be adequately and simply modelled. The method also eases the extraction of the required feedforward filters from the matrix equation; under ideal conditions the feedback paths cancel exactly and the feedforward filters are given by the matrix equation $\boldsymbol{C}^{-1}\boldsymbol{E}$. This is a familiar expression; it is the matrix form of the one dimensional situation consisting of a single detector and single speaker controlling the field at a single monitor position.

Consider such a single channel controller (figure 3). It can be realised simply by just a pair of electronic filters; a feedback path modelling the acoustic feedback and a feedforward filter of transfer function E/C.

Multiple Detectors, Multiple Secondary Sources

Figure 2 indicates that a multichannel controller can be readily realised by repeatedly using a number of the filter pairs used in the single channel control system. The implementation of a single channel controller therefore tests the basic unit of a multichannel system. However, it can be seen that the number of filter pairs needed is equal to the square of the number of channels (where each channel consists of a detector-speaker pair) thereby limiting the number of channels that can be implemented practically.

PRACTICAL IMPLEMENTATION OF A SINGLE CHANNEL CONTROL SYSTEM

This section contains a description of the experimental results obtained from an implementation of a single channel broadband control system partially attenuating the reverberant field inside an enclosure. The filter pair required was implemented as two 128 point FIR filters realised using a Texas instruments TMS32020 microprocessor housed in a Ferranti PC860XT personal computer. A method is needed whereby the coefficients for the (digital) control filters described above can be obtained. The practical method used in these experiments consisted of a series of acoustical measurements on the control system. The same hardware was used both to record the various frequency responses of the system from which the digital filters were derived and also to implement the controller. This ensured an easy means whereby the electronic filter compensated for its own imperfections and ensured that the sampling rates used for the various measurements and for the subsequent filter implementation were all the same.

A suitable test enclosure (0.5 x 0.6 x 0.7m), practical apparatus and test conditions were configured to produce a situation in which a control system could be successful (figure 3). The first two modes of the enclosure had modal frequencies at about 240 and 290 Hz. Therefore the working range of the system was conditioned to be up to 350 Hz (determined by the cut off of the low pass filters at the entrance to and exit from the digital system). The sampling rate used was 1 kHz. Measurements were recorded by exciting the system with a swept frequency sine wave output from the digital system and capturing the response on the same digital system. The following measurements were recorded: a transient swept sine signal (x_2) was used to excite the noise source loudspeaker and responses captured at the detector microphone monitor (y_{10}) and the monitor microphone (y_{30}); the signal y_{10} was used to excite the control speaker and the response captured at the monitor microphone (y_{32}); the transient swept sine signal was used to excite the control speaker and the response was captured at the detector microphone (y_{12}).

The feedback filter was derived from a deconvolution of the signals x_2 and y_{12}. The deconvolution was achieved with a least squared error FIR fit in the time domain. The feedforward filter was derived by deconvolving the signals y_{30} andy_{32}.

RESULTS

The results of the practical implementation of the control system operating in the enclosure are shown in figure 4. The noise source was driven with a pseudo-random signal from a Hewlett packard spectrum analyser. The signal from the monitor microphone was connected to the spectrum analyser to record the transfer function between the noise source signal and the signal at the monitor. The response with and without the control system in operation is shown. The control system was stable and attenuated the field to the same extent months after the system had been set up and the digital filters had been derived, demonstrating substantial stability over time.

CONCLUSIONS

A simple topology for the controllers for a multichannel control system has been presented. The method demonstrates how any multichannel controller can be realised by repeatedly using a number of the same type of filter pairs used in the single channel control system. A single channel broadband digital control system consisting of a single detector microphone and a single speaker attenuating the field at a single monitor position has been implemented. The active system successfully attenuated the first two modes of the reverberant field inside an enclosure.

ACKNOWLEDGMENTS

The work reported here was supported by an SERC CASE studentship in cooperation with Plessey Research (Caswell).

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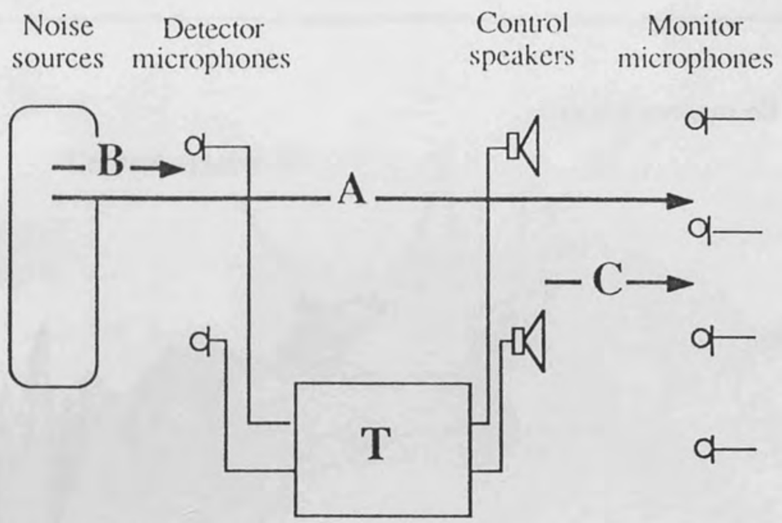

Figure 1. Two channel active noise control system

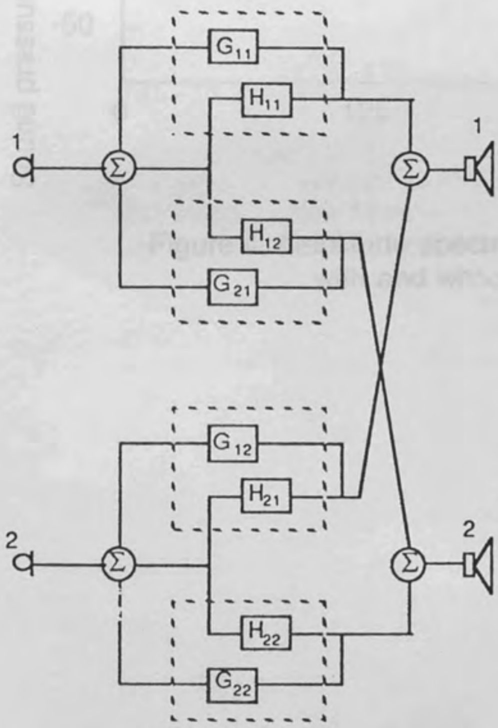

Figure 2. Two channel controller

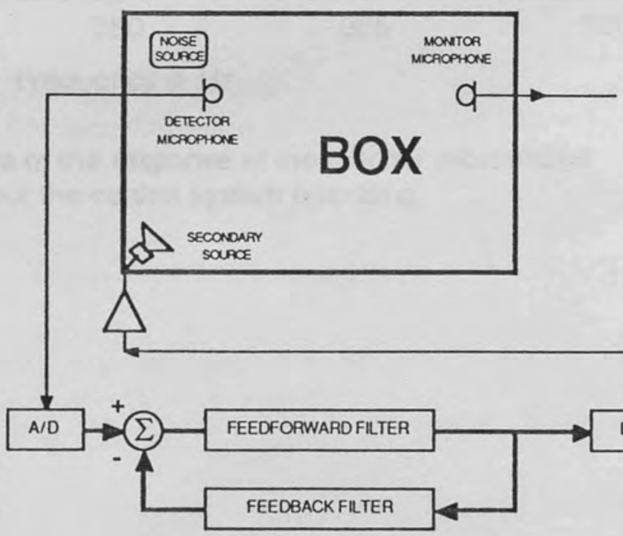

Figure 3. Single channel active noise control system in an enclosure

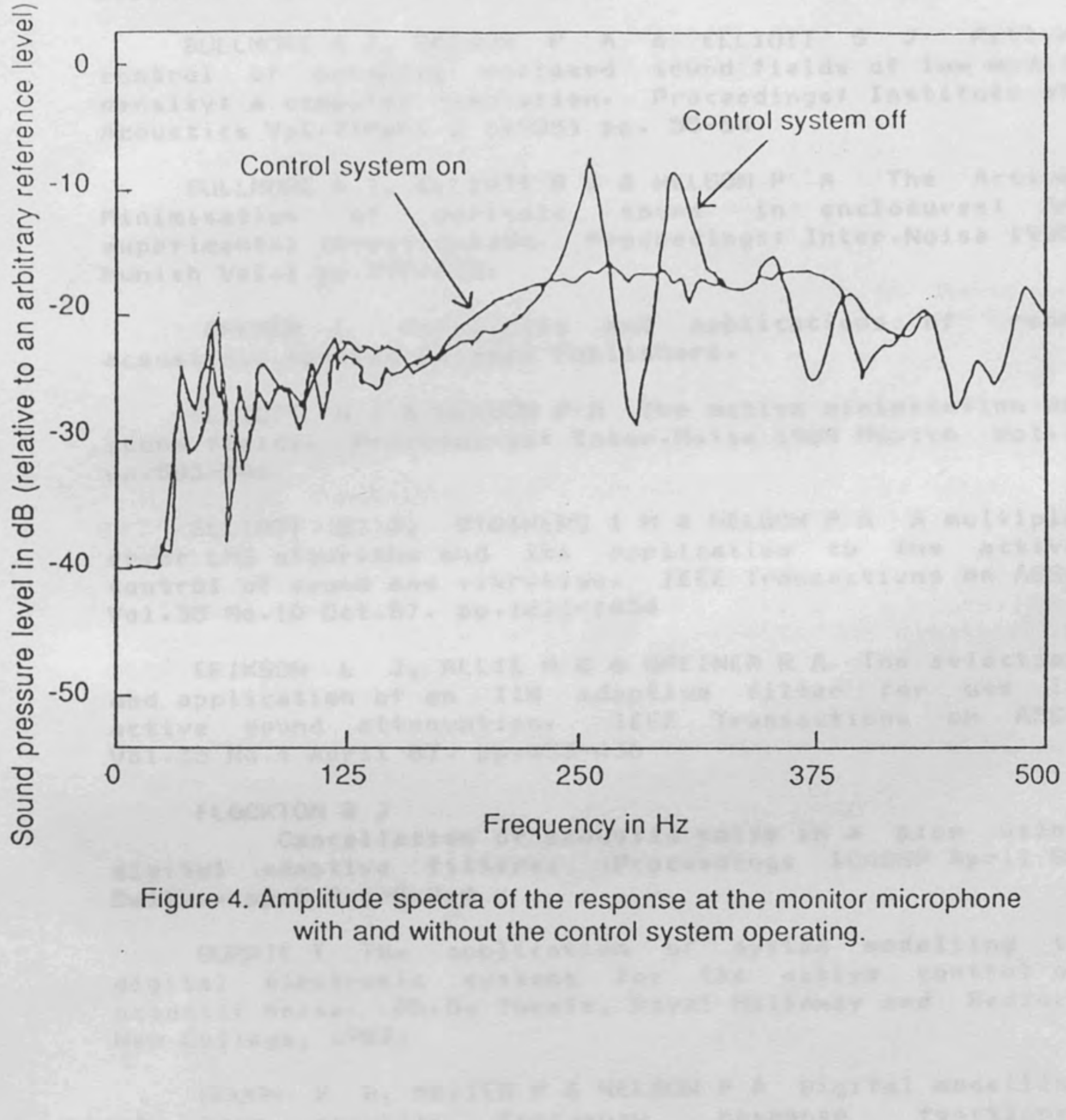

Figure 4. Amplitude spectra of the response at the monitor microphone with and without the control system operating.

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