Modeling of and acoustic feedback cancellation in hearing instruments

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**MODELING OF AND
ACOUSTIC FEEDBACK CANCELLATION IN
HEARING INSTRUMENTS**

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SCHOOL OF ELECTRONIC & ELECTRICAL ENGINEERING

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To all members in my family,
for their love and never-ending encouragement

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Abstract

Acoustic feedback is one of the challenges in hearing instrument design where the amount of acoustic gain of the hearing instrument is limited before the onset of acoustic feedback. When acoustic feedback occurs, the hearing instrument is effectively inoperable. Adaptive feedback cancellation algorithms embodied in digital hearing instruments is the method often adopted by manufacturers to address the acoustic feedback problem.

In this research program, an analogue behind-the-ear (BTE) hearing instrument is constructed and the known two-port modeling technique is applied to obtain the model of each component, and thereafter the open loop transfer function (OLTF). By means of the Nyquist Stability Criterion, the collective contribution and effect of each component to the acoustic feedback problem can be determined. The derived model of the hearing instrument is verified experimentally. The two-port model is found to be cumbersome, and to be largely inappropriate for the modeling of digital instruments because it is unable to accommodate non-linear algorithms.

Two digital hearing instruments are constructed and their models are obtained by means of the system identification method. One model involves only the feedback path, appropriate for simulations involving non-continuous feedback cancellation algorithms. The other model is the entire hearing instrument, appropriate for simulations involving continuous feedback cancellation algorithms. For the simplicity of modeling, we propose

to divide the entire instrument into five sub-blocks and then recombine them. The derived models are verified experimentally.

Finally the adoption of a one-step-ahead method is proposed for a non-continuous feedback cancellation system to address the deterioration of traditional non-continuous feedback cancellation algorithms due to environmental noise. We show, by means of a simulation, that our proposed adoption of the one-step-ahead method is able to mitigate some of deterioration due to the environment noise.

Glossary

ADC:	Analogue-to-Digital Converter
AGC:	Automatic Gain Control
AR:	Auto Regression
ARMAX:	Autoregressive moving average with exogenous input
BTE:	Behind-The-Ear
CIC:	Completely-In-Canal
DAC:	Digital-to-Analogue Converter
DSP:	Digital Signal Processing/Processor
FIR:	Finite Impulse Response
FFT:	Fast Fourier Transform
FXLMS:	Filtered-X LMS
HTL:	hearing threshold level
IIR:	Infinite Impulse Response
ITE:	In-The-Ear
LMS:	Least Mean Square
LS:	Least Square
MSG:	Maximum Stable Gain
NAL-NL:	National Acoustic Laboratories – Nonlinear
OLTF:	Open Loop Transfer Function
UCL:	Uncomfortable Level

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Chapter 1 Introduction

1.1 Hearing Impairment

Alongside the sensation of sight, hearing is probably the most valued of the human sensory systems. Unfortunately, hearing impairment (hearing loss) in present-day society has now become one of the most prevalent impairments [1]. It has been estimated that approximately 6%-10% of the world's population suffers from hearing impairment, ranging from the mild to the profound.

In general, hearing impairment may be classified into two forms, namely sensorineural and conductive [1, 2]. In both forms of hearing impairment, the hearing impaired suffers from reduced speech intelligibility compared to the normal hearing. It has been estimated that more than 90% of hearing impaired are of the former form or a combination of both forms. In most cases of sensorineural hearing impairment, it is untreatable by drugs or surgical procedures.

Hearing impairment is typically quantified by an audiogram and an example is depicted in Fig. 1.1. In Fig. 1.1, the abscissa is frequency (Hz), usually within the speech range from 125Hz to 8kHz. The ordinate, on the other hand, is the hearing level measured in dB HTL (Hearing Threshold Level), usually within the range of -10dB to 120dB. Hearing threshold is defined as the softest sound level that the person under evaluation can hear. The cross indicates the hearing threshold of the left ear and the circle, the hearing threshold of the right ear. Two persons' hearing thresholds are depicted in Fig. 1.1. The hearing thresholds of a normally hearing person are shown with

solid lines while the hearing thresholds of a person with a hearing impairment are depicted with dashed lines.

Fig. 1.1 Audiogram of a general sensorineural hearing loss

Although hearing impairment can occur at all frequencies within the speech range, most of the hearing impaired suffer from severe hearing loss in the high frequencies. Physiologically, this is attributed to the increased mechanical stresses to the hair cells in the basal end of the cochlear. To compensate for the hearing impairment, thereby improve the speech intelligibility of the hearing impaired, audiologists and/or audiometrists routinely prescribe hearing instruments (hearing aids) with acoustical gains according to prescription formulas, for example the NAL-NL (National Acoustic Laboratories - Nonlinear) formula. There is a number of hearing instrument gain

prescription formulas and in general, the acoustical gain of the hearing instruments is generally adjusted to between one-third to one-half of the hearing thresholds.

In addition to the acoustical gains, hearing instrument characteristics do account for the reduced dynamic range of the hearing impaired. The dynamic range at a given frequency is defined as the range of sound intensities from the softest sound audible (the threshold) to the loudest sound (the Uncomfortable Level, UCL) tolerated by the hearing impaired. For instance, if the acoustical gain at 1kHz is 40 dB, the output of the hearing instrument becomes excessive when the input is >70 dB – the output is hence > 110 dB, assuming this intensity is at the UCL. To accommodate the reduced dynamic range of the hearing impaired, compression is usually applied where the gain of the hearing instrument is reduced beyond a certain level, the compression knee. For example (assuming input compression), if the compression knee is set at 40dB and the compression ratio set at 2:1, the gain below an input of 40 dB is 40dB and beyond 40 dB, the gain is 20 dB. In this example, the UCL is reached when the input is now 90 dB instead of 70 dB formerly. To ensure that the output of the hearing instrument does not exceed the UCL, the maximum may be limited by several means, including peak clipping and/or various compression algorithms.

1.2 Hearing Instruments

The hearing instrument is essentially an acoustical amplifying device and as previously discussed, it is the most widely used device to enhance the intelligibility of speech of the hearing impaired [4]. Compared with other hearing assistive devices, the hearing instrument has two major advantages:

- a. The benefits obtained in terms of increased speech intelligibility exceeds that achieved by most of the other prosthetic devices, and
- b. The treatment can be easily applied and terminated at any time, and the cost of the device is relatively low.

The first commercially available hearing instruments were probably introduced at the beginning of 20th century [4~6]. They comprised a microphone, a carbon receiver, and a bulky battery. Amplification was achieved through the transduction process from acoustical to electrical signal in the carbon microphone. As a result, these early hearing instruments were of poor quality and provided limited gain. The quality and gain were improved with the invention of electron tubes, the first active components, and the development of practical electronic amplifiers. At the end of the 1950s, the first Behind-the-Ear (BTE) hearing instrument appeared. This new device has an enormous advantage over all the earlier types as it could be worn entirely on the ear. In the 1980s, two new styles of hearing instruments that could be worn inside the ear appeared. These instruments are commonly known as In-the-Ear (ITE) hearing instrument, worn within the concha of the pinna, and Completely-In-Canal (CIC) hearing instrument, worn within the ear canal. Apart from the aesthetic consideration, the main advantages of these small hearing instruments include small size, natural placement of the microphone, thereby enabling directional hearing (localization).

Fig.1.2 illustrates the block diagram of a typical electronic hearing instrument. It consists of five major parts: the microphone, pre-conditioner, signal processing unit, post-conditioner, and receiver (sub-miniature loudspeaker transducer). According to the type

of signal processing unit, electronic hearing instruments can be classified into two forms, namely analogue hearing instrument and digital hearing instrument [7, 8]. In both forms, the microphone picks up the incoming acoustical signal and converts it to a weak analogue electrical signal. In analogue hearing instruments, the pre-conditioner amplifies the weak signal to a higher signal level. In digital ones, the pre-conditioner not only amplifies the input signal but also converts the analogue signal into a digital form by means of an Analog-to-Digital Converter (ADC). The signal processing unit processes the amplified electrical audio signal with the objective of providing acoustical gain largely commensurate with the hearing impairment according to a fitting prescription formula (usually by means of filters), thereby improving the intelligibility of the hearing impaired. The post-conditioner of analogue hearing instruments provides the final amplification and sufficient power to drive the receiver while that of digital hearing instruments converts the digital signal back to analogue signal first prior to final amplification.

Fig. 1.2 Block diagram of electronic hearing instruments

From the hearing instrument user's viewpoint, it is highly desirable that the hearing instrument have the attributes given below. Some of these parameters can be qualified according to the usual hearing instrument measurement standards, the ANSI S3.22-1996

(or the IEC 118 standards) in more definitive parameters. From a user's perspective, the desirable attributes include:

- a. Acceptable sound quality,
- b. Acoustics of the amplified sound 'matches' that of the user's hearing impairment,
- c. Good speech intelligibility in all situations, including noisy environments,
- d. Perception of loudness,
- e. No acoustic feedback (whistling), and
- f. Easy management

Digital hearing instruments, as they embody a digital signal processor (DSP) with advanced signal processing algorithms, provide more capabilities/functionalities than their analogue counterparts in many aspects, including:

- a. Increased flexibility of the magnitude response to match the fitting prescription,
- b. Increased flexibility and control of compression processing,
- c. Implementation of adaptive feedback cancellation algorithms,
- d. Implementation of noise reduction algorithms,
- e. Implementation of speech enhancement algorithms, and
- f. Improved localization with directional microphone arrays

Despite the promise of higher ‘intelligence’ of current-art digital hearing instruments (as opposed to analogue instruments) from the confluence of the advances of digital technologies, there remain several unresolved challenges and problems [9]. Of the various challenges, acoustic feedback remains a difficult problem in a number of cases, particularly where the hearing instrument gain is high (e.g. >40 dB) and where there is an acoustical feedback path between the microphone and the receiver (see next section later). The acoustic feedback problem is particularly problematic because it limits the maximum usable gain of the instrument and degrades the overall system response, thereby reducing the speech intelligibility of the hearing impaired. In most cases, the occurrence of acoustic feedback causes the hearing instrument to be inoperable. In this dissertation, our effort in constructing and modeling a digital hearing instrument will be described and thereafter our effort in proposing an algorithm to reduce the possibility of acoustic feedback in digital hearing instruments will be introduced in detail.

1.3 Acoustical Feedback

In this section, the mechanism of acoustic feedback is described followed by a review of the research efforts reported in literature pertaining to acoustic feedback. Fig. 1.3 depicts an In-the-Ear (ITE) hearing instrument worn in a human ear. The amplified sound output from the receiver in the ear canal leaks out between the ear mould and concha of the hearing instrument user (and via the vent, if present) and re-pickup by the microphone in the ITE. In such a mode, a close loop is formed and the possibility of acoustic feedback exists.

Fig. 1.3 An ITE hearing instrument worn on a human ear

The mechanism for acoustic feedback is well established in literature and is largely based on the theory of oscillation in closed loop systems- the Nyquist Stability Criteria. Consider a simplistic model of the hearing instrument depicted in Fig. 1.4 where $H(j\omega)$ represents the transfer function of the feedback path and $G(j\omega)$ represents the intended transfer function (from the microphone of the instrument to the receiver output at the ear mould or ear shell) of the hearing instrument. In an analogue hearing instrument system, the transfer function of the feedback path includes the post-amplifier, receiver, acoustical feedback path, microphone, and pre-amplifier. In a digital hearing instrument system, on the other hand, the feedback path includes the digital-to-analogue converter (DAC), receiver, acoustic feedback path, microphone, and the analogue-to-digital converter (ADC). The details of these models will be exemplified in Chapters 2 and 3 later.

Fig. 1.4 Block diagram of amplification and feedback in a hearing instrument system

The transfer function of the entire hearing instrument system is

$$T(j\omega) = \frac{G(j\omega)}{1 - G(j\omega)H(j\omega)} \quad - (1.1)$$

where the loop gain $G(j\omega)H(j\omega)$ is also known as the open-loop transfer function (OLTF). It is well established that when

$$|G(j\omega)H(j\omega)| \geq 1 \quad - (1.1(a))$$

and

$$\angle G(j\omega)H(j\omega) = m360^\circ \quad - (1.1(b))$$

where m is an integer, the system is unstable, hence resulting in oscillation and physically, acoustic feedback occurs. For completeness, note that when $|G(j\omega)H(j\omega)|$ is slightly less than unity but $\angle G(j\omega)H(j\omega) = m360^\circ$, sub-oscillation occurs – although there is no oscillation, the speech quality is severely deteriorated by near occurrence of acoustic feedback.

Note that analogue hearing instruments and digital hearing instruments share largely the same components in the feedback path, such as the receiver, the acoustical path, and the microphone. This is one of the motivations to model the OLTF of an analogue hearing instrument albeit its limited application to contemporary digital hearing instruments. This topic will be discussed in more detail in Chapter 2.

1.3.1 Variations in the Feedback Path

A well-known observation of the feedback path is that it is affected by the changes in the environment [10]. For example, the acoustical feedback path of a hearing instrument placed in-situ would change when a telephone handset is moved close to the hearing instrument, when a hand is brought up to the hearing instrument to adjust the volume control, or when the position of the instrument shifts in the ear when the user eats or talks, etc.

For example, in a simple experiment, we measured the variations of the feedback path of a BTE hearing instrument by placing a 1.44-inch floppy disk in 3 different settings: when the floppy disk is (i) removed (or infinitely placed), (ii) 10mm and (iii) 2mm away from the microphone of the hearing instrument. The magnitude and phase responses of the feedback path are depicted in Fig.1.5 (a) and (b) respectively.

In Fig.1.5(a), the line labeled (i) is the magnitude response of the feedback path when the floppy disk is absent while the lines labeled (ii) and (iii) are magnitude responses when the floppy disk is 10mm and 2mm away from the hearing instrument microphone respectively. Note that the magnitude response is increased by the placement of the floppy disk closer to the hearing instrument, especially at the frequencies range beyond 2

kHz. In view of the Nyquist Stability Criteria and that $G(j\omega)H(j\omega)$ is increased, the probability of oscillation is increased. These observations are also well established in literature [11].

Fig. 1.5 Variations in the feedback path when 1.44-inch floppy disk is placed in 3 different settings:

- (i) Removed, (ii) 10mm, and (iii) 2mm away from hearing instrument microphone :
 (a) Magnitude response (b) Phase response

1.4 Modeling Hearing instruments

There has been much research into the methodologies and algorithms to mitigate acoustical feedback in hearing instruments. These research involve the modeling of hearing instruments and the modeling provides details of the various components

embodied and their contribution or effect to the overall hearing instrument model. This section presents a literature review of the reported modeling methods to model hearing instrument, first for analogue instruments and thereafter for digital instruments.

1.4.1 Modeling the Open Loop Transfer Function of an Analogue Hearing Instrument

The development of analogue hearing instruments chronologically preceded the digital hearing instruments. As the sophistication of signal processing of analogue hearing instrument is limited, acoustic feedback is often reduced by physical means, such as reducing the size of the vent, or improving the fit of the ear mould, etc. The modeling of the components in an analogue instrument is useful as it provides insight into the contribution of the individual components and hence, the possibility of replacing certain components to reduce the possibility of the occurrence of acoustic feedback. This is one object of this research program. As mentioned earlier, analogue hearing instruments and digital hearing instruments share most components of the feedback path, the modeling work of an analogue hearing instrument is also helpful to guide the construction of a digital hearing instrument.

In Section 1.3, it was discussed that the Nyquist Stability Criteria involves both the feedback path $H(j\omega)$ and the feed-forward path $G(j\omega)$, hence the OLTF = $G(j\omega)H(j\omega)$. In view of this, the OLTF of hearing instrument systems have been investigated by many researchers, probably first by Egolf [12] in the 1970s pertaining to analogue hearing instruments.

The first modeling of the various components of a glass-type analogue hearing instrument initially involved the receiver [12] only, and later extended to include the external acoustic feedback path [13], microphone [14], amplifier-to-eardrum [15], and finally the entire OLTF [16]. Egolf's "two-port" model, which completely defines the dynamic behavior of the system, is based on four parameters which may be derived theoretically or from experiments. The "two-port" model can accommodate "loading" effects because it is based on energy flow, rather than signal flow, considerations. Based on Egolf's work, Kates [17] subsequently modeled an analogue ITE hearing instrument to predict the sound pressure near the tympanic membrane. During the course of modeling, Egolf concluded that the vent exiting through an in-situ hearing instrument could be treated mathematically as though it were a circular hole in an infinite plane baffle for purposes of computing the termination impedance seen by a plane pressure wave as it approached the vent exit from the inside of the ear mould. In other words, for this particular application, the human head was treated as a flat plate or sphere. The shortcoming of this simplification is that the effects of the pinna, head, and body are ignored, hence the somewhat inadequacy of the model.

To circumvent these limitations, Walsh [18] applied the well-known boundary element modeling method, which is based on a 3D image processing technique, to model the external human auditory system, including the pinna, head, body and the environment. Stinson *et al.* [19] subsequently applied the same boundary element modeling method to predict the effect of handset proximity on hearing instrument acoustic feedback.

The models derived from the "2-port" and the boundary element modeling method are essentially frequency-domain models and are useful to predict the oscillation frequency

and maximum stable gain of the hearing instrument according to the Nyquist Stability Criterion. To account for non-linear or time-variant algorithms, such as adaptive feedback cancellation system, embodied in advanced digital hearing instruments (see next section), a time-domain model is necessary.

Kates [20] attempted to derive a time-domain model in one specific example, by transforming the frequency-domain model to its time-domain equivalent by means of the well established impulse invariance or bilinear transformation methods. Although these methods could yield a time-domain model with a roughly matched magnitude response, it does not account for the phase response. This is a major shortcoming as the phase is a necessary parameter to predict the oscillation frequency and the maximum stable gain before the onset of acoustic feedback – refer to the Nyquist Stability Criteria in Eqn. (1.1).

As the aim of the modeling an analogue hearing instrument is to investigate the causes of unstable acoustic feedback and the approximate contribution of each component to OLTF, we will adopt Egolf's method to model the external acoustical path (the effects of pinna, head, and body are, however, ignored) and model the OLTF in frequency domain only. This will be detailed in Chapter 2 later. A more precise time-domain model of a digital hearing instrument, which considers the pinna, head, and body effects as well as the phase response, by means of the system identification method will be reviewed in the next section and applied in Chapter 3.

1.4.2 Modeling the Feedback Path of a Digital Hearing Instrument with System Identification Method

The modeling of the feedback path of a digital hearing instrument is different from that of an analogue hearing instrument. This is largely because of the non-linear (e.g. adaptive) algorithms embodied in the former, the contribution of each component to the OLTF can be determined, leading to the complete frequency-domain model and an example will be delineated in Chapter 2.

Several researchers have obtained time-domain models by various means. Bustamante [11] used a sequence of ‘clicks’ as the stimulus and measured the responses. The impulse response, which may be considered as a time-domain model of the system, was calculated by averaging the response. Kaelin *et al.* [21] modeled the feedback path with a Finite-Impulse Response (FIR) filter with length N . The length N is generally determined by the peak-to-tail ratio of the impulse response. Kates [22] modeled the feedback path with a delay and a DC zero followed by an all-pole Infinite-Impulse Response (IIR) filter in series with a FIR filter. In fact, independent of the FIR structure or the IIR structure, they are all special cases of the system identification method [23, 24].

1.4.2.1 System Identification Method

The system identification method is a well established modeling method and when compared to the physical modeling method (such as the “two-port” method [12]), the system identification method considers the system under modeling as a “black box”. Nevertheless, undoubtedly a good appreciation of the physical attributes of the system

provides insight into the system. The basic progress of system identification [23] is illustrated in Fig. 1.6 and will now be delineated.

Fig. 1.6 The system identification flow-graph

a) Experiment Design

The design of an identification experiment includes several choices, including the specific signals to measure, when to measure them, and which signals to manipulate and how to manipulate them. It also includes practical aspects, such as how to condition the signals before they are sampled. In the following sections, we will later discuss the considerations for the design of the experiment in detail.

b) Data

The input-output data is sometimes recorded during a specifically designed identification experiment, where the user may determine which signals to measure, when to measure them and may also choose the specific input signals. The objective here in view of the experiment design is to make these choices so that the data become maximally informative, subject to constraints that may be at hand.

c) Model Set Choice

The possible model structures for linear discrete-time systems include the ARMAX models (autoregressive moving average with exogenous input). A special case of the ARMAX model is the modeling of the system as an FIR structure and the noise as white noise. We will later adopt this type structure for our model, see Chapter 3.

d) Criterion of Fit

Suppose a set of candidate models has been selected and it is parameterized as a model structure using a parameter vector w . The search for the best model within the set then

becomes a problem of determining or estimating w . There are many different ways of organizing such a search and also different views on what one should search for. In our problem, we choose the minimum Least Square (LS) criterion.

e) Model Calculation

After the model structure and fit criterion have been determined, the experiment can commence the input-output data obtained and the model determined by computer-generated algorithms.

f) Model Validation

The model should be validated for verification of the model. This also serves to avoid an incorrect experiment design, collection of incorrect data, or incorrect structure choice for the model. If the determined model does not match the desired performance, the entire procedure would be reiterated to obtain a better model. In Chapter 3, we will construct an ITE digital hearing instrument and model its feedback path by following the system identification flow-graph depicted in Fig. 1.6.

1.4.3 Modeling the Entire Digital Hearing Instrument

To simulate the continuous feedback cancellation algorithms, modeling only the feedback path model is insufficient. In the simulation of a continuous feedback cancellation algorithm, the proper parameter set in the feed-forward path, such as the gain and the compression ratio, etc, affects the performance of feedback cancellation algorithms significantly [35]. In the following paragraph, we will introduce the general method to determine the gain in the hearing instrument to compensate hearing loss.

- a. The first step of the fitting prescription procedure is to determine the insertion gain, which is the difference between the gain with a hearing instrument worn and without.
- b. The second step of the fitting prescription procedure is to determine the gain of the hearing instrument by some criteria, such as NAL-NL. To determine the gain of the hearing instrument, we need the knowledge of the input path and the output path, which can be determined by our five block model.

In the modeling of the entire digital hearing instrument system, the modeling is simplified by modeling the entire hearing instrument as five independent sub-blocks: the input sub-block, output sub-block, feedback sub-block, vent sub-block, feed forward sub-block. We thereafter apply the system identification method to model each sub-block. The 5 sub-block models are finally recombined to model the entire hearing instrument system. In Chapter 3, the construction of a BTE digital hearing instrument will be described and how it will be modeled, and thereafter our experiments will be delineated to verify the model.

1.5 Feedback Reduction Methods

As explained briefly and alluded to in Section 1.4, the efforts of modeling is to facilitate the development and analytically evaluate the efficiency of algorithms for the reduction of acoustic feedback in hearing instruments. This acoustic feedback problem has been well recognized since the early days of audio systems, in particular public address systems.

Many approaches have been reported in literature to reduce acoustic feedback in hearing instruments. Generally, the feedback reduction methods can be classified into several types and we will now briefly discuss them in turn.

a. Gain Reduction Method

The most direct method to reduce acoustic feedback is to reduce the overall and/or high-frequency gain of the hearing instrument. In terms of Eqn.(1.1), $G(j\omega)$ is decreased, thereby reducing the possibility of feedback according to the Nyquist Stability Criterion.

This method is objectionable in typical hearing instrument fittings because the acoustic gain of the hearing instrument may not meet the prescribed acoustical gain and the hearing instrument user would hence not obtain the full benefits of amplification. Put simply, the gain of the hearing instrument being lower than prescription would usually reduce the speech intelligibility, and somewhat defeats the intention/purpose of the hearing instrument.

b. (Adaptive) Notch Filter Methods

The notch filter method [26], either fixed or adaptive, is essentially an improved gain reduction method – attenuation of the gain of the hearing instrument only at the frequencies where oscillation intends to occur. This is the case because $G(j\omega)$ in Eqn. (1.1) is a function of frequency. This method of feedback suppression for hearing instruments is adopted from its use in the audio industry for public address systems [28].

This approach is usually realized differently in different types of hearing instruments. In simple hearing instruments where there is only one channel (the speech audio

spectrum is not divided into separate bands), a notch filter is adjusted such that the stop-band of the filter is set to the frequency where oscillation intends to occur – effectively reducing the gain of the hearing instrument at the frequency.

In some more sophisticated multi-channel hearing instruments, the speech signal is separated into a number of bands and signals in each band are processed individually. In these instruments, the gain of the channel(s) where oscillation tends to occur is reduced or in some cases, a notch filter is placed within that particular channel.

Overall, the notch filter approach is superior to the simplistic approach of reducing the overall gain of the hearing instrument (hence the gain reduction in the complete speech spectrum) because in the former, only one section of the speech spectrum is reduced. In other words, the notch filter potentially offers increased speech intelligibility over the gain reduction method.

There are nonetheless several problems with the notch filter approach. First, as oscillation may occur at more than one frequency, more than one notch filter is required. Second, the phase of the filter tends to change most rapidly near the band edge of the filter and this may affect the adjacent frequency regions where the oscillation may occur. In other words, the stop-band of the notch filter may need to be dynamically readjusted to determine where the notch filter is most effective. Third, the stop-band of the notch filter presents a stop-band in part of the entire speech spectrum, possibly resulting in reduced speech intelligibility of the hearing instrument user. In Chapter 4, we will describe the adaptive notch filter system we developed with Simulink.

c. Phase Shifting Method

The phase shifting method involves the incorporation of a phase compensator in the hearing instrument amplifier ($G(j\omega)$) in Eqn. 1.1) which dynamically changes the phase of the OLTF by means of the insertion of a periodic variable delay [29-31].

This method can significantly increase the insertion gain before oscillation occurs at low frequencies. However its efficacy decreases rapidly at higher frequencies. In our experience, the phase shifting method offers little advantage, if any, as oscillations tend to occur at relatively high frequencies.

d. Frequency Shifting Method

Frequency shifting has been used rather extensively to reduce acoustical feedback in public address systems [32] and this classical method has been reported for the same in several hearing instruments [33]. The concept behind this method is the frequency shifting of the input signal to either a slightly lower or higher frequency before it reaches the hearing instrument receiver in amplified form. Frequency shifting method is based on the fact that it takes some time for oscillation to occur even when the oscillation conditions are met. However, the added stable gain achieved by this method is limited.

e. Fixed Feedback Cancellation Method

Levitt *et al.* [34] proposed a feedback cancellation system with a time-invariant (fixed) cancellation filter set when fitting the hearing instrument. In this method, the feedback signal is estimated and subtracted from the microphone input signal. Compared with the approaches mentioned above, this method is of, at least theoretically, of separating the

acoustic feedback signal from the desired signal. However, the time-invariant cancellation filter is limited because it cannot adjust to changes in the acoustic environment. In view of this, the more recent feedback cancellation methods are dynamic or adaptive.

f. Adaptive Feedback Cancellation

The adaptive feedback cancellation is dynamic by adjusting its coefficients to track the changes in the feedback path. Generally the adaptive feedback canceller can be classified into two forms: non-continuous and continuous. As its name implies, the continuous feedback canceller updates its coefficients continuously without interrupting the normal operation of the hearing instrument. On the other hand, the non-continuous feedback canceller updates its coefficients only when oscillation or a quiet environment is detected. If oscillation or quiet is detected, the normal operation of the hearing instrument is then interrupted with the insertion of white noise as the reference signal to the adaptive filter. The coefficients are then updated and upon completion, the hearing instrument resumes its normal operation.

Consider now the modeling of a hearing instrument and the model of a general feedback canceller. This model will serve to facilitate the review of the continuous and non-continuous feedback cancellation systems in the next two sections.

Fig. 1.7 depicts the model of an entire digital hearing instrument including the basic structure of an adaptive feedback cancellation system. $F(z)$ represents the transfer function of the feedback path, which is a cascade of the DAC, receiver, acoustic feedback path, microphone, and the ADC (Fig. 1.8). $W(z)$ represents the feedback canceller, either

fixed or adaptive. $G(z)$ represents the transfer function of the feed-forward path, which is the processing done in the DSP chip. The signal $u(n)$ is the output of the digital signal processor comprising the usual processed signal (during normal operation), white noise, $w(n)$, or both. The input signal of the hearing instrument, $s(n)$, is the combination of two signals, one from the feedback signal $f(n)$, the other from the desired signal $x(n)$. The estimated feedback signal, $\hat{f}(n)$, is subtracted from $s(n)$ and the resulting $e(n)$ is used to estimate the desired signal $x(n)$. In the ideal case, the estimated feedback signal $\hat{f}(n)$ matches the real feedback signal $f(n)$ with no bias (see later for definition of bias in the next section) and $e(n)$ can predict $x(n)$ precisely. The desired signal $x(n)$ can be modeled by passing a white noise signal $v(n)$ through a monic and inversely stable filter $H(z)$.

Fig. 1.7 Basic structure of feedback cancellation system

Fig.1.8 The structure of the feedback path $F(z)$

On the basis of the above model, a review of the continuous and non-continuous feedback cancellation algorithms will now be presented.

1.5.1 Continuous Feedback Cancellation System

Of the two feedback cancellation systems, the continuous and non-continuous forms, the continuous feedback canceller has attracted the most interest in literature. This is largely because it is ‘proactive’ in the sense that the functionality of the hearing instrument is uninterrupted independent of whether oscillation is detected or otherwise. The non-continuous feedback canceller, on the other hand, interrupts the normal operation of the hearing instrument after oscillation or when silence is detected.

The first proposed continuous feedback canceller was probably first proposed by Bustamante [11] in 1989 where the feedback path was modeled as a cascade of a delay and a finite impulse response (FIR) filter. The coefficients of the FIR filter were adjusted according to the well-known normalized least mean square (LMS) algorithm. The receiver input signal, $u(n)$, is the output of the feed-forward path $G(z)$. If the desired input signal $x(n)$ is spectrally colored, such as speech signal or pure tone signal, some correction will exist between the desired signal and the reference signal of the adaptive filter, $u(n)$, depending on the features of $x(n)$ and $G(z)$, and this causes a biased estimation $W(z)$ [35, 36]; bias is the mean difference between the ideal feedback path model and the model actually computed. The filter will cancel the feedback signal but because of the inadvertent and undesired bias, the filter will undesirably cancel the feedback signal but because of the inadvertent and undesired bias, the filter will undesirably cancel part of the

input signal as well. For completeness, it is probably interesting to note that Siqueira *et al.* [35] and Hellgren *et al.* [36] subsequently theoretically analyzed the bias problem but their analysis were limited to simple cases, such as adding a delay in the feed forward path. Due to the simplicity of the cases, their applications were applicable only to limited real applications.

As mentioned before, the bias problem exists because the correlation between the input signal and the reference signal. Arguably, the most direct way to reduce the bias problem in continuous feedback cancellation systems is to insert a signal to de-correlate the operations in the feed-forward path. Delay in the feed-forward path [37] is the simplest de-correlation operation and this is particularly attractive as it requires only a small amount of added computation and the output signal is only slightly degraded; note that in some systems, a delay is intrinsic in the feed-forward path, especially in those systems which process input signal frame by frame [21]. It is important, however, to appreciate that there is a limit to the amount of delay that can be inserted for two reasons. First, many hearing impaired make use of visual cues to improve their speech intelligibility and these visual cues should be synchronized with the audio signals provided by the hearing instrument. It is well established in psycho-acoustics that if the audio signal and the visual cues differ by 10-15 ms, the amount/degree of information provided by the visual cues diminishes. Second, this concerns cases of unilateral hearing impairment where the hearing impaired has one normal ear and a hearing impaired ear. It has been established that if the latency between the input and output of the hearing instrument is >12 ms, the unilateral hearing impaired donning the hearing instrument will

not benefit from this hearing instrument as much as that compared to an instrument whose latency is shorter.

In an attempt to reduce the delay, Park *et al.* [38] and Chankawee *et al.* [39] proposed a time-varying all-pass filter method to de-correlate the input signal $x(n)$ and reference signal $u(n)$. To achieve better de-correlation effects, a higher order filter is required in the feed-forward path and this undesirably results in more delay and the undesired latency between the impaired ear and the good ear in the unilateral hearing impaired case. This method, nonetheless, introduces a shorter delay than the previous methods of directly introducing a delay.

An alternative popular method to de-correlate the input signal and reference signal [11, 40] is to add delay in the feedback cancellation path. Siqueira *et al.* [35] analyzed the effect of inserting delay in the feedback path from a theoretical perspective. He showed that although the bias caused by the correlation of the desired input signal $x(n)$ and the reference signal $u(n)$ may be reduced by the delay inserted, new undesirable bias results from the replacement of useful head samples of the impulse response of feedback path with zeros (due to the delay).

Hellgren [41] and Kates [42] subsequently proposed a method, named Filtered-X LMS (FXLMS) method, to reduce the bias problem as well as increasing the convergence rate by whitening the reference signal. In the FXLMS method, the desired input signal $x(n)$ is modeled by passing a white noise signal $v(n)$ through a fixed filter $H(z)$. FXLMS corresponds to the system identification with an output error model. The feedback path is then identified under the assumption that the input signal to the system is white. The

basic structure of FXLMS method is illustrated in Fig. 1.9, where $L(z)$ represents the inverse of $H(z)$ and parameters $s(n)$, $u(n)$, $f(n)$, $W(z)$ represent the same meaning with those in Fig.1.7. Widrow *et al.* [43] provided a theoretical analysis on FXLMS method where it was assumed that the auto-regression filter $H(z)$ is fixed. However, this assumption is not practical because speech signal is not be modeled well by a fixed filter.

Fig. 1.9 Filtered-X System

Chi [44] subsequently improved the general FXLMS method by band-limiting the receiver input $u(n)$ and error signal $e(n)$. The basis of this improvement is due to the fact that the energy of the input signal $x(n)$, such as speech signal, focuses on low-frequency region below 1 kHz and oscillation in hearing instruments tends to occur at high-frequency normally above 2 kHz. This band-limited FXLMS method also employs a fixed whitening filter.

Due the impractical assumption assumed by a number of continuous feedback cancellers, the performance of continuous system in a practical hearing instrument may be somewhat degraded, although Maxwell *et al.* [40] and Greenberg *et al.* [45] claim (and as adopted by most of hearing instrument manufacturers) that continuous feedback cancellers have better performance and are preferred by hearing instrument users.

In addition, as the reference signal of the adaptive filter in the continuous system is highly dependent on the input signal, some of these systems may be unstable. Kates [46] attempted to overcome this disadvantage by constraining the feedback canceller in specified regions.

1.5.2 Non-Continuous Feedback Cancellation System

The non-continuous feedback cancellation system is less attractive than the continuous systems largely because of its “reactive” feature, that is, the system does not avoid oscillation before it occurs. Another disadvantage of the non-continuous system is that the white noise that is frequently inserted as a reference signal is annoying especially since the normal operation of the hearing instrument is interrupted frequently. Nonetheless, this disadvantage can be partially mitigated by inserting white noise probes of short durations (<50 ms) which are nearly inaudible.

In spite of the disadvantages of non-continuous systems, they remain of interest to the research community because of its stability relative to the continuous-time system and because it does not suffer from the bias problem. In Chapter 4 of this dissertation, we will describe our proposed non-continuous system.

The first non-continuous feedback cancellation can probably be attributed to Kates [27], where the coefficients of the feedback canceller were updated only when oscillation is detected. Maxwell [40] extended the non-continuous system to update the coefficients when oscillation or quiet is detected. In these reported work, including others, the performance of non-continuous feedback cancellers is severely deteriorated when oscillation is detected in a noisy environment. A special and common case is that

oscillation occurs when the hearing instrument wearer is talking with other people. Note that to the feedback canceller, the speech signal is a noise signal. It is hence desirable to find a solution to precisely estimate the feedback path in noisy environments for the non-continuous system to be practical as noisy environment are prevalent in everyday living. In Chapter 4, we propose and apply one-step-ahead prediction method to a non-continuous system to reduce the noisy-coefficient problem.

1.6 Objectives

The primary objectives of the research program are to model hearing instruments and to propose an algorithm to mitigate the acoustical feedback problem in hearing instruments, thereby allowing an increase in the gain of the hearing instrument before the onset of acoustic feedback. The specific objectives are to:

- a. Investigate the contribution of each component in a hearing instrument to the acoustic feedback problem,
- b. Construct hearing instrument systems and thereafter model them so that they may be sued for simulating feedback cancellation algorithms, and
- c. Develop a non-continuous feedback cancellation algorithm that features better performance than a traditional algorithm in a noisy environment.

1.7 Major Contributions of the Thesis

The major contributions of the thesis pertain to modeling hearing instruments and the proposal of a feedback cancellation algorithm that has improved performance over a reported algorithm in a noisy environment. In summary, the contributions are the:

- a. Construction of an analogue hearing instrument and modeling of each component in the open loop transfer function. With these models, we can derive the frequency domain model of the entire model of the open loop transfer function. With the Nyquist Stability Criteria, we can then determine the component in the open loop feedback path, such as the vent length and diameter, transducers, etc., that may be replaced/adjusted to mitigate the acoustical feedback problem.
- b. Application of the system identification method to model the (i) open-loop transfer function of the digital hearing instrument, and (ii) entire digital hearing instrument. The entire hearing instrument is divided into five sub-blocks and the system identification method is extended to model not only the feedback path (one of the sub-blocks), but also the other sub-blocks such as the input and output sub-blocks. This model is useful to determine the parameter set of the forward path, such as gain and compression rate, for simulating non-continuous and continuous feedback cancellation algorithms.
- c. Proposal of a non-continuous feedback cancellation algorithm to mitigate the possibility of acoustic feedback, thereby increasing the gain of the hearing instrument before the onset of acoustic feedback. The novelty is the

application of a one-step-ahead method to improve its performance in noisy environments. The improved efficacy is shown in a limited example.

1.8 Organization of the Thesis

In this chapter, an overview of hearing impairment, hearing instruments, the mechanism of acoustic feedback in hearing instruments, and possible solutions to reduce acoustical feedback reported in literature has been presented. The objectives and major contributions of the research program have also been presented. In Chapter 2, the construction of an analogue hearing instrument, the modeling of each component in the open loop transfer function (OLTF) based on reported methods and finally the entire model to achieve the model of the OLTF will be discussed. In chapter 3, the construction of a digital hearing instrument, the model the OLTF, and the extension of the system identification method to model the entire hearing instrument system is described. In Chapter 4, the one-step-ahead prediction method is proposed and applied to a non-continuous adaptive feedback cancellation system. Finally, in Chapter 5, conclusions are drawn and the future work is presented.

Chapter 2 Modeling the Open Loop Transfer Function of an Analogue Hearing Instrument

2.1 Introduction

As discussed in Chapter 1, one of the primary objectives of the research program is to investigate the acoustical feedback problem in hearing instruments and the modeling of hearing instruments facilitates this. The Nyquist Stability Criteria was also discussed where the criterion that when the magnitude response of the OLTF is greater than unity and the phase response of the OLTF is multiple of 2π , oscillation will occur at that frequency. If each component in the OLTF can be modeled and the model of the OLTF derived, the oscillation may be determined. From the hearing instrument design point of view, it may be possible to select different components, such as another microphone and/or receiver with different characteristics, and this would alter the characteristics of the OLTF. A selection of appropriate components can yield a hearing instrument that is less prone to oscillation and the modeling of these components and eventually the OLTF facilitates this selection.

In Chapter 1, it was explained that Egolf [12] first employed the "two-port" model of the OLTF of a glass-type analogue hearing instrument. In this chapter, we will apply Egolf's method to model an analogue BTE hearing instrument, and will derive the mathematical expression of the OLTF. We would thereafter provide the expressions to determine the parameters in OLTF model. The parameters are obtained from different sources, including manufacturers' manuals, theoretical expressions, and experiments.

Modeling the Open Loop Transfer Function of an Analogue Hearing Instrument

Finally, we will illustrate the OLTf model achieved by cascading the model of each component in the open loop and verify the model experimentally. For completeness, note that much of the work in this chapter is based on an emulation of Egolf's [12] work but adopted to our constructed BTE hearing instrument.

2.2 Analogue BTE Hearing Instrument

In this section, we will describe our constructed analogue BTE hearing instrument. Fig. 2.1 illustrates the instrument mounted on the pinna of a B&K (Bruel & Kjaer) head and torso simulator 4128C (manikin). The sound waves of pressure P at some distant point in space are transformed by transfer function I to sound pressure P_I at the input point of the Knowles EA-1843 microphone (9), which is characterized by transfer function M . The microphone transduces sound pressure signal P_I to an electrical signal that undergoes amplification in the amplifier (12), which is characterized by transfer function A . Note the amplifier here is a linear amplifier; a non-linear amplifier cannot be modeled by the two-load method. The amplified electrical signal drives the Knowles BK-1604 receiver (1), which is characterized by transfer function R . The output sound pressure P_R from the receiver is transformed to sound pressure P_T at the point close to the microphone in the ear simulator 4158C by a serial connection of waveguides, namely the connector (2), hook (3), plastic tubes (4, 5, 6), adaptor (10), and coupler (11). The respective transfer functions are characterized by T_2 , T_3 , T_4 , T_5 , T_6 , T_{10} , and T_{11} , respectively.

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Fig. 2.1 Analog BTE hearing instrument mounted on the pinna of a manikin

- 1: Receiver 2: Connector 3: Hook 4, 5, 6: Plastic tubes 7: Vent 8: External path
9: Microphone 10: Adaptor 11: Coupler 12: Amplifier

Some sound in the ear simulator leaks via the vent (7) in the ear-mould, denoted by transfer function T_7 . The sound pressure P_O at the internal end (in coupler) of the vent becomes P_V at the vent outlet. The sound pressure P_V then undergoes further transformation to P_F via the external acoustic feedback path, which is denoted by transfer function F . Upon reaching the microphone (of the hearing instrument) inlet port, sound pressure P_F combines with sound pressure P_I . Such a summation process occurs naturally

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when two sound pressure waves - in this case P_I and P_F - arrive simultaneously at the same point in space.

2.3 Two-Port Model of the OLTF

We will now develop the two-port model expression of the OLTF based on the two-port method, using the model and nomenclature in section 2.2. Fig. 2.2 depicts the signal flow block diagram of the analogue hearing instrument system, where T_{23456} represents the cascaded transfer function of the connector (2), hook (3), and plastic tubes (4, 5, and 6) while T_{10_11} represents the cascaded transfer function of the adaptor (10) and coupler (11). This is a simplified diagram composed of "one-port" blocks, which does not account for the "loading" phenomenon in some components of the hearing instrument.

Fig. 2.2 Block diagram of a typical BTE hearing aid based on signal-flow concept

With reference to Fig. 2.2, the OLTF may be analytically obtained by (a) cutting the feedback path just prior to the point at which P_F joins the summer, as highlighted by a dashed line in Fig. 2.2, and (b) deriving an expression for the ratio of P_F/P_S . The desired ratio, obtained by inspection of Fig. 2.2, is

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$$P_F/P_S = GH \quad - (2.1)$$

where

$$G = \text{MART}_{23456} \quad - (2.2)$$

and

$$H = T_7 F \quad - (2.3)$$

where G is the transfer function of the forward path and H is the transfer function of the feedback path. Note that there are many different definitions for the forward path and feedback path in literature. For example, the amplifier may be deemed as the only component in the forward path and the receiver, acoustical feedback path, and the microphone deemed as components that constitute the feedback path.

In order to account for the fact that the "loading" effect occurs between some adjacent components of a hearing instrument, the diagram of Fig. 2.2 is altered to a "two-port" network [12], which is illustrated in Fig. 2.3. Here two-port blocks replace their one-port counterparts in Fig. 2.2. The transfer function of each component in Fig. 2.2 has been replaced by four parameters, A , B , C , and D . The two-port block in Fig. 2.3 can accommodate the "loading" effects because it is based on energy flow, rather than signal flow, considerations. The energy-flow designation is derived from the fact that the product of any pair of block input or block output variables has units of power. Note that in the event that loading does not occur, the two-port block may then be reduced to a one-port block.

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Fig. 2.3 Two-port network diagram of the analogue hearing instrument system

Fig. 2.4 Revised Two-port diagram of Fig. 2.3

To facilitate the development of an expression for the OLTF, the circuit of Fig. 2.3 is redrawn as Fig. 2.4 after cutting the feedback path and making several changes. The act of cutting the feedback path effectively open circuits the output terminals of the block. This is equivalent to attaching an infinite impedance load $Z \approx \infty$ across these terminals after cutting the path. Such action does not adversely affect computation of the OLTF because the particle velocity output u_F from the external path block was essentially zero (The input port of the hearing instrument microphone is very small/narrow, i.e. the input impedance is very high) before the cut was made. In Fig. 2.4, the original circuit, minus the incident path and summer, has been straightened out. Also, the input impedance to the ear canal impedance Z_E and vent radiation impedance Z_v have both been converted to equivalent two-port blocks.

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By combining all the individual blocks, Fig. 2.4 can be simplified to Fig. 2.5 and the expression for the circuit is given as

$$\begin{Bmatrix} P_s \\ u_s \end{Bmatrix} = \begin{bmatrix} A_{GH} & B_{GH} \\ C_{GH} & D_{GH} \end{bmatrix} \begin{Bmatrix} P_F \\ u_F \end{Bmatrix} \quad - (2.4)$$

where

$$\begin{bmatrix} A_{GH} & B_{GH} \\ C_{GH} & D_{GH} \end{bmatrix} = \begin{bmatrix} A_M & B_M \end{bmatrix} \begin{bmatrix} A_A & B_A \end{bmatrix} \begin{bmatrix} A_R & B_R \end{bmatrix} \begin{bmatrix} A_{2-6} & B_{2-6} \end{bmatrix} \begin{bmatrix} 1 & 0 \\ 1/Z_E & 1 \end{bmatrix} \begin{bmatrix} A_7 & B_7 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ 1/Z_v & 1 \end{bmatrix} \begin{bmatrix} A_F & B_F \\ C_F & D_F \end{bmatrix} \quad - (2.5)$$

Fig. 2.5 Reduction of Fig. 2.4 by matrix multiplication

The A , B , C , and D entries in Eqns. (2-4) and (2-5) are cascade parameters of the blocks in Fig. 2.4 which are represented by the subscripts of GH for the open loop system, M for the microphone, A for the amplifier, R for the receiver, F for the external path, and 2,3,4,5,6,7 for cylindrical acoustic waveguides.

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Using the two-port method, we have now derived an expression for the two-port model of the OLTF. To calculate the four entries in the model, we need to determine the two-port model of the microphone, amplifier, receiver, external feedback path, and cylindrical acoustic waveguides. At the same time, we will also need to determine the values of the input impedance to the ear canal Z_E and vent radiation impedance Z_V .

2.4 Entries in the analytical model

In this section, we will now determine/obtain the parameter entries of each individual entry in Eqn. (2-5), including those of microphone (M), amplifier (A), receiver (R), plastic tubes (2-6), vent (7), external path (F), and acoustical impedance Z_E and Z_V . Some of these parameters can be determined experimentally while others can be obtained analytically, or from manufacturers' datasheets.

2.4.1 External Acoustic Path

In this section, we will determine the relationship between the two-port network entry A_F , B_F , C_F , and D_F and the ratio of P_F/P_V . We will thereafter derive the analytical expression to determine P_F/P_V .

A general circuit equation for the external path block in Fig. 2.4 is given as:

$$\begin{Bmatrix} P_V \\ U_P \end{Bmatrix} = \begin{bmatrix} A_F & B_F \\ C_F & D_F \end{bmatrix} \begin{Bmatrix} P_F \\ u_F \end{Bmatrix} \quad - (2.6)$$

where P_V and U_P are the complex amplitudes of sound pressure and volume velocity at the output port of the ear mould vent, P_F and u_F are complex amplitudes of sound

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pressure and particle velocity arriving at the microphone input port via the external path; and A_F , B_F , C_F , and D_F are cascade parameters of the external path.

If the microphone draws little or no particle velocity u_s , then the particle velocities u_I and u_F delivered to the microphone via the incident and feedback paths would likewise be negligible. This means that

$$u_I = u_F \approx 0 \quad - (2.7)$$

A method that will account for the effects of Eqn. (2-7) is to set $B_F = D_F = 0$ in Eqn. (2-6), and the resulting equation can be further simplified by assuming that nearly all of the volume velocity U_V exiting the ear mould vent is absorbed by the vent radiation impedance Z_V . This means that very little, if any, of the volume velocity leaving the vent is actually delivered to the external path block in Fig. 2.3, resulting in $U_p \approx 0$. This can be accounted for by setting $C_F = 0$ in Eqn. (2-6) producing

$$\begin{Bmatrix} P_V \\ U_p \end{Bmatrix} = \begin{bmatrix} A_F & 0 \\ 0 & 0 \end{bmatrix} \begin{Bmatrix} P_F \\ u_F \end{Bmatrix} \quad - (2.8)$$

which is characteristic of a one-port model of the external path. The entry A_F can be determined by P_V/P_F .

We will now derive the analytical expression to determine the ratio P_V/P_F , based on that developed by Morse and Ingard and Egolf [12] where they concluded that the vent exiting through an in situ hearing instrument could be treated mathematically as though it were a circular hole in an infinite plane baffle for purposes of computing the termination impedance seen by a plane pressure wave as it approached the vent exit from inside the

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ear mould. In other words, for this particular application, the human head was treated as a flat plate. Note although it appears that it is common sense to treat a somewhat spherical object such as the human head as a flat plate, Egolf demonstrated that the plane model appears to yield somewhat better results than the sphere model. We will hence treat the human head as a plane baffle in this section.

In the following derivation, the side of the head is assumed to be infinite plane baffle through which the vent hole emerges, as shown in Fig. 2.6. The word "infinite" means that the baffle is very large compared to the longest wavelength of interest - the wavelength is much greater than the largest dimension of the head. The head model thus appears flat, is of infinite extent, and assumed to have no attached pinna, neck, or body.

Fig. 2.6 Setup for determining P_f/P_v for the case of a piston-shaped sound source located in an infinite plane baffle

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Assuming the vent exit of the in-situ ear mould may be treated as a piston-shaped sound source located in an infinite plane baffle as in Fig. 2.6, then the sound pressure P_F in the far-field of the source is given by Eqn. (2-9) of Kinsler *et al.*(1982) [56]:

$$P_F = jk\rho_0 c \left[\frac{U_v \pi r^2}{2\pi R} e^{-jkR} \left(\frac{2J_1(kr \sin \theta)}{kr \sin \theta} \right) \right] \quad - (2.9).$$

where

k is wavenumber (ω/c), m^{-1} ,
 ω is radian frequency ($2\pi f$), rad/s ,
 f is frequency, Hz ,
 ρ_0 is air density, kg/m^3 ,
 c is sonic velocity, m/s ,
 U_v is velocity amplitude of the piston-shaped source, m/s ,
 r is radius of the vent, m ,
 R is distance from vent exit to point of interest, m ,
 θ is angle from vent centerline to point of interest, deg ,
 J_1 is a Bessel function of the first kind of order one, dimensionless, and
 j is $\sqrt{-1}$

Eqn. (2.9) gives the relationship between P_F and U_v . As the velocity amplitude of the piston-shaped source U_v is identically the particle velocity amplitude at the vent exit, sound-pressure amplitude P_v at the vent exit is related to U_v by:

$$U_v = P_v / Z \quad - (2.10)$$

where Z is radiation impedance of a piston-shaped sound source set in an infinite plane baffle and is given by [57]

$$Z = \rho_0 c \left[\left(1 - 2 \frac{J_1(2kr)}{(2kr)} \right) + j \left(2 \frac{H_1(2kr)}{(2kr)} \right) \right] \quad - (2.11)$$

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Where J_1 is Bessel function of the first kind of order one, dimensionless and H_1 is a Struve function of order one, dimensionless. Substitution of Eqns. (2.10) and (2.11) into Eqn. (2.9) and canceling the redundant numerator and denominator terms yields

$$\frac{P_F}{P_V} = \left\{ \frac{kr^2 e^{j(\pi/2 - kr)}}{2R \left[\left(1 - 2 \frac{H_1(2kr)}{(2kr)} \right) + j \left(2 \frac{H_1(2kr)}{(2kr)} \right) \right]} \right\} \left(\frac{2J_1(kr \sin \theta)}{(kr \sin \theta)} \right) \quad - (2.12)$$

From Eqn. (2.12) and using Matlab, we simulate the magnitude and phase response of the acoustic path and this is illustrated in Fig. 2.7.

Fig. 2.7 Frequency response of external acoustical path

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The top figure in Fig. 2.7 is the magnitude response and shows that there is about 25 dB attenuation of sound-wave pressure at all the frequencies. The bottom figure is the phase response and shows that the external acoustic path has a linear phase.

2.4.2 Vent radiation impedance Z_v

The acoustic vent radiation impedance Z_v appearing can be approximated as the impedance of a piston of a radius r equal to that of the vent mounted in a plane baffle and described analytically by Eqn. (2.13) [57] below. The frequency response of Z_v is depicted in Fig. 2.8.

$$Z_v = \rho_0 c \left[\left(1 - 2 \frac{J_1(2kr)}{(2kr)} \right) + j \left(2 \frac{H_1(2kr)}{(2kr)} \right) \right] \quad - (2.13)$$

Fig. 2.8 The frequency response of Z_v

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From Fig. 2.8, it can be seen that the acoustic impedance decreases as the frequency increases. The bottom figure is phase response, which has no special meaning.

2.4.3 Cylindrical Waveguides

The connector (2), hook (3), plastic tubes (4, 5, and 6), vent (7), adaptor (8) and coupler (9) are considered as cylindrical tubes. The expressions to determine the cascaded parameters of the damped acoustical plane-wave propagation through cylindrical tubes may be represented analytically by the expressions

$$\begin{aligned} A_n &= \cosh \Gamma_n L_n \\ B_n &= Z_n \sinh \Gamma_n L_n \\ C_n &= Z_n^{-1} \sinh \Gamma_n L_n \\ D_n &= \cosh \Gamma_n L_n \end{aligned} \quad - (2.14)$$

where L_n is length of the tube, Γ_n is the propagation operator, and Z_n , is the characteristic impedance. The expressions to determine Γ_n and Z_n are given by [58]

$$\Gamma_n = j \frac{w}{c} \left(\frac{1 + 2(\gamma - 1) \left[\frac{J_1(\alpha a_n)}{\alpha a_n J_0(\alpha a_n)} \right]}{1 - 2J_1(\beta a_n) / \beta a_n J_0(\beta a_n)} \right)^{1/2} \quad - (2.15)$$

$$Z_n = \frac{\rho_0 c}{\pi a_n^2} \left[\left(1 - \frac{2J_1(\beta a_n)}{\beta a_n J_0(\beta a_n)} \right) \times \left(1 + 2(\gamma - 1) \frac{J_1(\alpha a_n)}{\alpha a_n J_0(\alpha a_n)} \right) \right]^{1/2} \quad - (2.16)$$

where

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$$\begin{aligned}\alpha &= (-j\omega\rho\sigma / \mu)^{1/2} \\ \beta &= (-j\omega\rho / \mu)^{1/2}\end{aligned}\quad - (2.17)$$

The parameters of each cylindrical waveguide according to Eqn. (2.14) are simulated and illustrated in Figs. 2.9-2.16.

Fig. 2.9 Real and imaginary part of entries of connector

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Fig. 2.10 Real and imaginary part of entries of hook

Fig. 2.11 Real and imaginary part of entries of plastic tube I

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Fig. 2.12 Real and imaginary part of entries of plastic tube II

Fig. 2.13 Real and imaginary part of entries of plastic tube III

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Fig. 2.14 Real and imaginary part of entries of adaptor

Fig. 2.15 Real and imaginary part of entries of coupler

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Fig. 2.16 Real and imaginary part of entries of vent

The common feature of Figs. 2.9 - 2.16 is that the C entry is very small and the D entry is nearly unit, which means the (volume) velocity of the particles in a tube is largely unaffected by the sound pressure.

2.4.4 Input impedance to the ear canal Z_E

The expression to determine the converted ear impedance Z_E is derived as follows:

$$\begin{Bmatrix} P_o \\ U_o \end{Bmatrix} = \begin{bmatrix} A_{10-11} & B_{10-11} \\ C_{10-11} & D_{10-11} \end{bmatrix} \begin{Bmatrix} P_T \\ U_T \end{Bmatrix} \quad - (2.18)$$

$$Z_E = \frac{P_o}{U_o} = \frac{A_{10-11}P_T + B_{10-11}U_T}{C_{10-11}P_T + D_{10-11}U_T} = \frac{A_{10-11}\left(\frac{P_T}{U_T}\right) + B_{10-11}}{C_{10-11}\left(\frac{P_T}{U_T}\right) + D_{10-11}} = \frac{A_{10-11}Z_T + B_{10-11}}{C_{10-11}Z_T + D_{10-11}} \quad - (2.19)$$

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where Z_T can be obtained from manufactures' datasheet.

The A_{10-11} , B_{10-11} , C_{10-11} , and D_{10-11} can be obtained by:

$$\begin{bmatrix} A_{10-11} & B_{10-11} \\ C_{10-11} & D_{10-11} \end{bmatrix} = \begin{bmatrix} A_{10} & B_{10} \\ C_{10} & D_{10} \end{bmatrix} \begin{bmatrix} A_{11} & B_{11} \\ C_{11} & D_{11} \end{bmatrix} \quad - (2.20)$$

where the entries of adaptor (10) and coupler (11) have been obtained in last section. The calculated Z_E is illustrated in Fig. 2.17. In Fig. 2.17, there is clear a peak around 2.5 kHz.

Fig. 2.17 The frequency response of converted ear acoustic impedance

2.4.5 Microphone

As delineated by Egolf [12], we similarly employ the "two-load" method to model the microphone EA-1843 by attaching it to two different loads. We first derive expressions

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for A_M , B_M , C_M and D_M and thereafter design experiments to achieve the parameters needed. In our experiment, the two electrical loads are applied sequentially to the transducer output terminals, as depicted in Fig. 2.18 (a) and (b). The two loads in Fig. 2.18 are both resistive: R_L having a large value of electrical resistance and R_S having a small electrical resistance value. The values of R_L and R_S are selected so as to span the magnitude of impedance loads normally seen by the microphone.

Fig. 2.18 The sequential application of two different resistive loads, R_L and R_S to the two-port model of hearing aid microphone

For the two-port model of Fig. 2.18(a), circuit equations are written in matrix form and the load equation is given as

$$\begin{Bmatrix} P \\ u \end{Bmatrix} = \begin{bmatrix} A_M & B_M \\ C_M & D_M \end{bmatrix} \begin{Bmatrix} E \\ I \end{Bmatrix} \quad - (2.21)$$

and

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$$E = R_L \times I \quad - (2.22)$$

Similar equations for the two-port model of Fig. 2.18(b) are written as

$$\begin{Bmatrix} P' \\ u' \end{Bmatrix} = \begin{bmatrix} A_M & B_M \\ C_M & D_M \end{bmatrix} \begin{Bmatrix} E' \\ I' \end{Bmatrix} \quad - (2.23)$$

and

$$E' = R_S \times I' \quad - (2.24)$$

where the primes have been added to indicate changes in the original variables that are reflective of the changes from R_L to R_S .

As previously mentioned, the sub-miniature hearing aid microphone used in the Knowles EA-1843 has very small input ports. Consequently, the input impedance of such microphones is extremely high—much greater than that of most sound fields—and they draw very little particle velocity from such fields. For this microphone, it may hence be assumed that

$$u = u' \approx 0 \quad - (2.25)$$

which results in

$$C_M = u / E |_{I=0} = u' / E' |_{I'=0} \approx 0 \quad - (2.26)$$

and

$$D_M = u / I |_{E=0} = u' / I' |_{E'=0} \approx 0 \quad - (2.27)$$

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Substitution of Eqns. (2.26) and (2.27) to Eqns. (2.21) and (2.23) produces

$$\begin{Bmatrix} P \\ u \end{Bmatrix} = \begin{bmatrix} A_M & B_M \\ 0 & 0 \end{bmatrix} \begin{Bmatrix} E \\ I \end{Bmatrix} \quad - (2.28)$$

and

$$\begin{Bmatrix} P' \\ u' \end{Bmatrix} = \begin{bmatrix} A_M & B_M \\ 0 & 0 \end{bmatrix} \begin{Bmatrix} E' \\ I' \end{Bmatrix} \quad - (2.29)$$

Substitution of Eqns. (2.22) and (2.24) into the first of Eqns. (2.28) and (2.29) for $/$ and F yields

$$P / E = A_M + B_M / R_L \quad - (2.30)$$

and

$$P' / E' = A_M + B_M / R_S \quad - (2.31)$$

which can be solved simultaneously for A_M and B_M to yield

$$A_M = [(P / E)R_L - (P' / E')R_S] / (R_L - R_S) \quad - (2.32)$$

and

$$B_M = R_L R_S (P' / E' - P / E) / (R_L - R_S) \quad - (2.33)$$

Eqns. (2.32) and (2.33) are mathematical relationships for the cascade parameters of the two-port model in Fig. 2.18.

The two ratios P/E and P'/E' need to be determined experimentally. The experimental setup for measuring these quantities on a sub-miniature hearing instrument microphone is illustrated in Fig. 2.19. With reference to Fig. 2.19, the measuring microphone (8) is

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located in close proximity to the hearing instrument microphone (2) so as to monitor sound pressure in the field directly adjacent to the hearing aid microphone. The electrical resistor R_n (4) represents both R_L and R_s , which are applied separately and sequentially. It should also be noted that a decoupling capacitor C_d (3) has been inserted between the hearing aid microphone and the load resistor R_n . This capacitor was included to prevent the load resistor R_n from affecting the dc bias point of the amplifier—an FET source follower—which is an integral part of the particular hearing instrument microphone being tested.

Fig. 2.19 Experimental setup for measuring transfer functions T_{PE} and T'_{PE}

The spectrum analyzer (6) in Fig. 2.19 utilizes the familiar fast Fourier transform algorithm to compute the following transfer functions which include the ratios P/E and P'/E' :

$$T_{PE} = K_M K_A P / (K_B E) \quad - (2.34)$$

and

$$T'_{PE} = K_M K_A P' / (K_B E') \quad - (2.35)$$

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where K_M is the transfer function of the measuring microphone and K_A and K_B are the transfer functions of the conditioning amplifiers. Frequency-dependant values of K_M are available in the microphone manufacturers' datasheet and the amplifier transfer functions K_A and K_B can be obtained experimentally with the spectrum analyzer. Substitution of Eqns. (2.34) and (2.35) into Eqns.(2.32) and (2.33) for the ratios P/E and P'/E' produces

$$A_M = \frac{K_B(R_L T_{PE} - R_S T'_{PE})}{K_M K_A (R_L - R_S)} \quad -(2.36)$$

and

$$B_M = \frac{R_L R_S K_B (T'_{PE} - T_{PE})}{K_M K_A (R_L - R_S)} \quad -(2.37)$$

Matlab is used to determine the A_M and B_M cascade parameters of a miniature hearing aid microphone via Eqns. (2.36) and (2.37). The input data to this program required by Eqns. (2.36) and (2.37) include:

- a. The instrument transfer functions K_M , K_A and K_B , where K_M is from the manufacturers' manual and K_A and K_B can be obtained from the spectrum analyzer. The frequency response of K_M , K_A and K_B are illustrated in Figs. 2.20, 2.21 and 2.22, respectively,
 - b. The presented values of R_L and R_S ($R_L = 9.90 \text{ k } \Omega$ and $R_S = 990 \text{ } \Omega$), and
 - c. The transfer functions T_{PE} and T'_{PE} , which is illustrated in Fig. 2.23 and 2.24.
- Computer generated values of A_M and B_M for the Knowles EA-1843 miniature hearing aid microphone are plotted in Fig. 2.25.

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Fig. 2.20 Frequency response of K_M

Fig. 2.21 Frequency response of K_A

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Fig. 2.22 Frequency response of K_B

Fig. 2.23 Ratio T_{PE} with large load (mark as T_{pet})

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Fig. 2.24 Ratio T_{PE} with small load (mark as T'_{pes})

Fig. 2.25 Calculated A_M and B_M in real and imaginary terms

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Fig. 2.26 Open circuit sensitivity determined experimentally and from manufacturer's datasheet

Fig. 2.26 shows that the open circuit sensitivity from our experiments agrees with that from manufacturer's datasheet for frequencies below 8 kHz. The difference is less than 2dB.

2.4.6 Amplifier and Receiver

Using the same method to model the microphone, we can obtain the cascaded entries of the (linear) amplifier and receiver. Instead of modeling the amplifier and receiver separately and then cascades them together, the "two-load" method can be applied to the amplifier-receiver system as a "black box" as illustrated in Fig. 2.27. In this section, the expressions to determine the model entries A_{AR} , B_{AR} , C_{AR} and D_{AR} will first be derived

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based on Egolf’s work [12] and the experiments to achieve the parameters will thereafter be designed.

Fig. 2.27 Two-port model of the cascaded amplifier and receiver

Fig. 2.28 the sequential application of two ((a) and (b)) different acoustical loads to the two-port model of amplifier-receiver

As in the case of the microphone modeling, two different loads are used to model amplifier-receiver. Note that the loads used to model amplifier-receiver- are acoustical loads, as illustrated in Figs. 2.28 (a) and (b). The entries of the low acoustical load are labeled A_L , B_L , C_L , and D_L , while the entries of the high acoustical load are labeled A_H , B_H , C_H , and D_H .

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At this juncture, we have obtained the entries A_M and B_M of the hearing instrument microphone. To verify the results we obtained, we compare them against the data published by the manufacturer. It is common practice for microphone manufactures to publish "open-circuit sensitivity" data for their products. Analytically, the open-circuit sensitivity of a microphone is simply the output voltage E divided by the input sound pressure P , with the electrical output terminals open circuited. In this case, there is no output current I , and the two-port network equivalent is to set $I=0$ in Eqn. (2.28). In other words,

$$E / P |_{I=0} = 1 / A_M \quad - (2.38)$$

These simulated values of the magnitude and phase of $1/A_M$ (computed via the inverse of Eqn. 2.36) for a Knowles EA-1843 sub-miniature hearing instrument microphone are given in Fig. 2.26. In Fig. 2.26 we also plot the magnitude of the open-circuit sensitivity of the same microphone, extracted from the manufacturer's published data sheet. Note that the manufacturer does not publish the phase of the open-circuit sensitivity, no phase data from the manufacturer is available.

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As before for the two-port model of Fig. 2.28(a), the circuit equations are written in matrix form and the load equation are given as

$$\begin{Bmatrix} E \\ I \end{Bmatrix} = \begin{bmatrix} A_{AR} & B_{AR} \\ C_{AR} & D_{AR} \end{bmatrix} \begin{Bmatrix} P \\ U \end{Bmatrix} \quad - (2.39)$$

$$\begin{Bmatrix} P \\ U \end{Bmatrix} = \begin{bmatrix} A_L & B_L \\ C_L & D_L \end{bmatrix} \begin{Bmatrix} P_M \\ U_M \end{Bmatrix} \quad - (2.40)$$

and

$$P_M = Z_M \times U_M \quad - (2.40)$$

Similar equations for the two-port model of Fig. 2.28(b) are written as

$$\begin{Bmatrix} E' \\ I' \end{Bmatrix} = \begin{bmatrix} A_{AR} & B_{AR} \\ C_{AR} & D_{AR} \end{bmatrix} \begin{Bmatrix} P' \\ U' \end{Bmatrix} \quad - (2.42)$$

$$\begin{Bmatrix} P' \\ U' \end{Bmatrix} = \begin{bmatrix} A_H & B_H \\ C_H & D_H \end{bmatrix} \begin{Bmatrix} P'_M \\ U'_M \end{Bmatrix} \quad - (2.43)$$

and

$$P'_M = Z_M \times U'_M \quad - (2.44)$$

where the primes have been added to indicate changes in the original variables that are reflective of the changes from the low acoustical load to the high acoustical load.

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Substitution of Eqns. (2.41) and (2.44) into the Eqns. (2.40) and (2.43) for U_M and $\mathcal{E}'_M$ yields

$$P = A_L P_M + B_L P_M / Z_M \quad - (2.45)$$

$$U = C_L P_M + D_L P_M / Z_M \quad - (2.46)$$

$$P' = A_H P'_M + B_H P'_M / Z_M \quad - (2.47)$$

$$U' = C_H P'_M + D_H P'_M / Z_M \quad - (2.48)$$

Substitution of Eqns. (2.45) - (2.48) into Eqns. (2.39) and (2.42) yields

$$E = (A_{AR} A_L + B_{AR} C_L) P_M + (A_{AR} B_L + B_{AR} D_L) \left(\frac{P_M}{Z_M} \right) \quad - (2.49)$$

$$I = (C_{AR} A_L + D_{AR} C_L) P_M + (C_{AR} B_L + D_{AR} D_L) \left(\frac{P_M}{Z_M} \right) \quad - (2.50)$$

$$E' = (A_{AR} A_H + B_{AR} C_H) P'_M + (A_{AR} B_H + B_{AR} D_H) \left(\frac{P'_M}{Z_M} \right) \quad - (2.51)$$

$$I' = (C_{AR} A_L + D_{AR} C_L) P'_M + (C_{AR} B_L + D_{AR} D_L) \left(\frac{P'_M}{Z_M} \right) \quad - (2.52)$$

The expressions of A_{AR} , B_{AR} , C_{AR} , and D_{AR} from Eqns. (2.49) - (2.52) can be determined if the other parameters in the equations can be obtained. To do this, we set up an experiment illustrated in Fig. 2.29.

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Fig. 2.29 Experimental setup to model the amplifier-receiver

The receiver (3) modeled is the Knowles BK-1604. The low acoustical load is the coupler which can be considered as a cylindrical tube. The high acoustical load is a cascade of a steel tube and the coupler. The entries of the low acoustical load and the high acoustical load determined analytically using the method discussed in section 2.4.3. The output from the microphone (5) is amplified by a conditioning amplifier (6) and analyzed by the spectrum analyzer (7).

With reference to Fig. 2.29, the expression can be derived [59]:

$$A_{AR} = \frac{Z_M (MS_A) [(C_H Z_M + D_H) / T_{PE} - (C_L Z_M + D_L) / T'_{PE}]}{(A_L Z_M + B_L)(C_H Z_M + D_H) - (A_H Z_M + B_H)(C_L Z_M + D_L)} \quad (2.53)$$

$$B_{AR} = \frac{Z_M (MS_A) [(A_L Z_M + B_L) / T'_{PE} - (A_H Z_M + B_H) / T_{PE}]}{(A_L Z_M + B_L)(C_H Z_M + D_H) - (A_H Z_M + B_H)(C_L Z_M + D_L)} \quad (2.54)$$

$$C_{AR} = \frac{Z_M (MS_A / 100) [(C_H Z_M + D_H) / T_{PI} - (C_L Z_M + D_L) / T'_{PI}]}{(A_L Z_M + B_L)(C_H Z_M + D_H) - (A_H Z_M + B_H)(C_L Z_M + D_L)} \quad (2.55)$$

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$$D_{AR} = \frac{Z_M (MS_A / 100) [(A_L Z_M + B_L) / T'_{PI} - (A_H Z_M + B_H) / T_{PI}]}{(A_L Z_M + B_L)(C_H Z_M + D_H) - (A_H Z_M + B_H)(C_L Z_M + D_L)} \quad - (2.56)$$

where

$$\begin{aligned} T_{PE} &= MS_A P_M / E_i \\ T_{PI} &= MS_A P_M / I_i \\ T'_{PE} &= MS_A P'_M / E'_i \\ T'_{PI} &= MS_A P'_M / I'_i \end{aligned} \quad - (2.57)$$

Z_M is the acoustical impedance of the diaphragm of the measuring microphone, M is the acoustic-electric transfer function of the same measuring microphone, and S_A is the transfer function of the microphone amplifier.

The parameters used in Eqns. (2.53) - (2.56) and the results of A_{AR} , B_{AR} , C_{AR} , and D_{AR} are illustrated in Figs. 2.30 -2.33.

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Fig. 2.30 two-port model of the low acoustical load

Fig. 2.31 two-port model of the high acoustical load

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Fig. 2.32 Frequency response of the microphone (*M*)

Fig. 2.33 the parameter entries of modeled amplifier-receiver

2.5 Modeling the OLTF

It has been shown that by carrying out the matrix multiplication in Eqn. (2.5), that the three zero entries in the "external path" matrix (Eqn 2.8) will result in $B_{GH}=C_{GH}=D_{GH}=0$. Hence Eqn. (2.5) can be simplified as:

$$\begin{Bmatrix} P_s \\ u_s \end{Bmatrix} = \begin{bmatrix} A_{GH} & 0 \\ 0 & 0 \end{bmatrix} \begin{Bmatrix} P_F \\ u_F \end{Bmatrix} \quad - (2.58)$$

where the first entry in the matrix on the right side of Eqn. (2.58) can be written as

$$GH = 1 / A_{GH} \quad - (2.59)$$

The expression given by Eqn. (2.59) is, in fact, the desired ratio P_F/P_s and therefore represents the OLTF of the block diagram in Fig. 2.3. The A_{GH} entry can be obtained by performing the matrix multiplication indicated in Eqn. (2.5).

With all the parameters now obtained, Matlab is used to determine the frequency response of the entire OLTF. This is depicted in Fig. 2.34. To verify the results, we experimentally measure the magnitude response of the OLTF at several frequencies and the results are also plotted in Fig. 2.34. The small difference may be attributed to the external acoustic feedback path, and to the fact that the pinna and the head effects have been ignored.

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Fig. 2.34 Frequency response of the open loop

From Fig. 2.34, it can be seen that there are two zero-phase frequencies: one is about 500 Hz, the other is about 3500 Hz. The magnitude response around 500 Hz, according to the model, is -7 dB and that round 3500 Hz is about -20 dB. In such case, if the gain of the hearing instrument is increased by 7 dB, oscillation will occur at 500 Hz. If we can replace an appropriate component in the OLTF, such as replacing the microphone by another one with the same magnitude response but different phase response, the zero-phase frequency may be shifted to a lower frequency, such as 300 Hz, where the magnitude response is about -18 dB. Hence the acoustical oscillation can be avoided until 17 dB gain is inserted.

As a general discussion in view of the objectives of work presented in this chapter, we find that the two-port modeling technique rather antiquated. In view of this, we do not

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advocate this method for modeling more advanced digital hearing instruments - it is also inappropriate/unable to model non-linear digital processing algorithms. In Chapter 3, we will adopt the system identification method for modeling hearing instruments.

2.6 Conclusions

The two-port method has been applied to model a specially constructed BTE hearing instrument placed in-situ on a manikin. The model of each component in the OLTF is determined experimentally, from theoretical analysis or from data sheets. The model of the entire OLTF is derived by cascading the model of each component, and has been verified against physical measurements. This two-port method has been found to be cumbersome for modeling hearing instruments and have limited application to digital hearing instruments.

Chapter 3 Modeling Digital Hearing Instruments

3.1 Introduction

In Chapter 1, we explained that the primary objectives of the research program are to study the acoustical feedback problem in hearing instruments and to efficiently suppress the acoustic feedback. In this research program, we will develop a feedback cancellation algorithm only for digital hearing instruments (instead of analogue because of the limited signal processing capability of digital hearing instruments). To simulate the algorithm, we will first construct a digital hearing instrument and thereafter model it. The modeling will be based on the system identification method delineated in Chapter 1. As already discussed in chapter 2, the two-port method used for the modeling of analogue hearing instruments is inappropriate largely because it is unable to accommodate non-linear algorithms embodied in many digital hearing instruments.

As introduced in Chapter 1, there are two types of feedback cancellation system: non-continuous and continuous. To simulate a non-continuous feedback cancellation system, the modeling of the feedback path is required. However, to simulate a continuous feedback cancellation system, the modeling of the entire hearing instrument system is required as the parameter set (such as gain etc.) used in the feed-forward path affects the performance of the continuous feedback cancellation system significantly.

In this chapter, we will first introduce the modeling of the feedback path of a digital hearing instrument. Thereafter, we will explain how the system identification method can be extended to model the entire digital hearing instrument system. Finally, we will

model the feedback path of an ITE hearing instrument and thereafter model an entire BTE hearing instrument. In both cases, they will be verified experimentally.

3.2 Modeling the Feedback Path of an ITE Hearing Instrument

As introduced in Chapter 1, several steps are involved in the application of the system identification method to model a feedback path. The first step is to design the experiment. The sketch of the experiment setup is illustrated in Fig. 3.1. Fig. 3.1(a) is a custom-made ITE hearing instrument that embodies only a microphone and a receiver. The ITE shell is placed in the concha of the ear simulator (2) of a manikin (1), depicted in Fig. 3.1(b). The microphone (5) and receiver (6) are connected to a DSP board (ADSP-2189) (10) via two custom-made signal conditioning circuits (8) and (9), which are cascade of a low-pass filter with 8 kHz cutoff frequency and an optional power amplifier. Part of the sound output from the receiver leaks back to the microphone through the vent in the ITE shell (not shown in Fig. 3.1(b)) and the wall of ear canal, which is indicated by a dashed line (4).

Fig. 3.1 (a) an ITE shell with microphone and receiver (b) Sketch of the experimental setup of the ITE hearing instrument placed in-situ

1. Manikin 2. Ear simulator 3. Pinna 4. Acoustical feedback path 5. Microphone of ITE
6. Receiver 7. Microphone to measure sound pressure level around ear membrane
8. signal conditioner 9. Signal conditioner 10. DSP board

In this experiment, the DSP is considered as the forward path ($G(j\omega)$) in Fig. 1.4) and the cascade of the other parts, including signal conditioning (9), receiver (6), acoustical feedback path (4), microphone (5), signal conditioning (8), ADC and DAC on the DSP board (10), is considered as the feedback path ($H(j\omega)$) in Fig. 1.4). To avoid the undesired effects of the environment noise, the experiment is conducted in a NC-15 sound proof room.

With reference to the system identification flow-graph in Fig. 1.6, the data refers to the stimulus signal and the response signal. The stimulus signal in this experiment is white noise generated by the DSP and the response signal in this experiment is the response of the feedback path, which is recorded by the DSP simultaneously. To avoid transients, the first few thousand samples of both the stimulus and the response are discarded.

In the model structure set, the first structure we choose for the feedback path is the simplest special case of the ARMAX [23] structure, which considers the feedback path as an FIR filter. The ARMAX model constitutes an important class of difference equations in the form:

$$A(z^{-1})y_k = z^{-d}B(z^{-1})u_k + C(z^{-1})w_k \quad - (3.1)$$

where d is a time domain delay, w_k is a white noise, u_k is the input signal, y_k is the output signal, and A, B, C are polynomials,

$$\begin{aligned} A(z^{-1}) &= 1 + a_1 z^{-1} + \dots + a_{n_A} z^{-n_A} \\ B(z^{-1}) &= b_0 + b_1 z^{-1} + \dots + b_{n_B} z^{-n_B} \\ C(z^{-1}) &= 1 + c_1 z^{-1} + \dots + c_{n_C} z^{-n_C} \end{aligned} \quad - (3.2)$$

with the unknown parameters

$$[a_1 \quad \dots \quad a_{n_A} \quad b_0 \quad \dots \quad b_{n_B} \quad c_1 \quad \dots \quad c_{n_C}]^T \quad - (3.3)$$

For an FIR filter structure, as a special case of ARMAX, $A(z^{-1}) = 1$ and $C(z^{-1}) = 1$ may be considered. Hence,

$$y_k = z^{-d} B(z^{-1})u_k + w_k \quad -(3.4)$$

We will later show that the FIR structure model achieved by our experiment is indeed sufficiently precise.

With the stimulus and response signal, according to the Least (LS) Square Fit Criteria, we can derive the time-domain model [23]. To ensure a sufficiently precise model, especially phase response, we choose FIR model instead of IIR model. We choose the length of FIR model to ensure that the peak-to-tail ratio of the impulse response to be larger than 40 dB (which is considered to be precise enough) [24]. The peak-to-tail ratio is defined as the largest value of the sample in the impulse response to the average absolute value of several samples in the tail of the impulse response. In this experiment, we find that the filter length up to 300 is sufficient to meet this requirement. For completeness, note that special attention is paid to ensure the stimulus signal is in the linear range of the system - this will be verified in Experiment 2 in Section 3.2.1 later. The experiment is repeated five times and we obtain five sets of data leading to five sets of impulse responses with the system identification algorithm. Thereafter, we determine the averaged impulse response and this is illustrated in Fig. 3.2.

Fig. 3.2 Impulse response of the feedback path

In Fig. 3.2, the x-axis is the tap number and the y-axis is the coefficient values. For convenience of illustration, we only display the first 100 taps in Fig. 3.2. The impulse response can be divided into three parts, the head, the body, and the tail. The value of head samples are very small due to the delay introduced by the external feedback path. The body samples, from the 10th tap to the 90th tap are the most important parts as they comprise most of the information. The tail samples are those taps beyond the 90th tap, which are almost all zero.

3.2.1 Experimental Verification

The procedure of the system identification method in Fig. 1.6 involves verification of the model and we design three experiments to do this. In the first experiment, we use a product, named Sound Check, an instrument to measure hearing instruments, to obtain

the frequency response of hearing instrument. This measured frequency response is thereafter compared against the frequency response obtained from the impulse response time-domain model in Fig. 3.2 by the well established FFT method.

In order to apply the model obtained here to simulate our feedback cancellation algorithm, the system under modeling should be linear. The second experiment is designed to verify this.

According to Nyquist Stability Criteria, an accurate model should be able to predict the oscillation frequency precisely. In the third experiment, we adjust the gain of the DSP of the hearing instrument until oscillations occur and measure the oscillation frequency to verify if the oscillation frequency predicted by our model is of the same frequency.

We will now delineate all three experiments in turn.

a. Experiment 1

Fig. 3.3 depicts the experimental setup for this experiment. Sound Check, via a sound card, generates a sweep signal as the stimulus and records the response simultaneously and the frequency response of the system is obtained. Note that this setup is different from our system identification method because the closed loop of hearing instrument system is 'cut' to allow the stimulus to be input and the response recorded. In this experiment, the closed-loop is cut at the dashed line illustrated in Fig. 3.1.

Fig. 3.3 Sweep signal method to measure the external feedback path

A sound card, named Card Deluxe (part of the Sound Check system), is inserted into a PC with the application software installed. A sweep signal from 600Hz to 6300Hz is injected into the signal conditioning circuit (9) and the output of the DSP board (10) is simultaneously recorded. Note here the DSP is programmed to be "run through" with no gain. The magnitude response and phase response of the feedback path achieved are depicted in Fig. 3.4.

Fig. 3.4 Frequency response of the external feedback path

As the sound card will introduce some intrinsic delay, we accommodate determining its delay by another experiment. To determine the intrinsic delay and thereafter accounting for the delay, we connect the output of sound card directly to the input of the sound card and the frequency response of the card itself is obtained. This is illustrated by the curves with asterisk (indicated as Deluxe) in Fig. 3.4. This delay cascaded with the frequency response of the feedback path is the actual measurement from the experiment illustrated in Fig. 3.3 and this is depicted as Deluxe-Ext in Fig. 3.4. The effect of the sound card is then subtracted from the cascaded frequency response and the frequency response of the feedback path can be derived. This is plotted in Fig. 3.4 as the line with triangles and indicated as Ext.

By means of the system identification method, the frequency response of the external feedback path is obtained by transforming the time-domain model (in Fig. 3.2) into and frequency domain by the well established FFT method. The two frequency responses are now compared and for ease of comparison, the two responses are overlaid together in Fig. 3.5.

From Fig. 3.5, it is noted that the two responses agree well between 1000Hz and 6000Hz, the frequency range where unstable oscillation would likely occur. Some differences are observed at low frequencies and this is probably due to the loading of the DAC. Nonetheless, the difference at low frequency is inconsequential because acoustic feedback rarely, if ever occurs <1000 Hz. This experiment verifies that the time domain model (impulse response) of the external feedback path derived from the system identification method is precise.

Fig. 3.5 Frequency response of the external feedback path obtained from the sweep signal method and the system identification method

b. Experiment 2

It is imperative that the system is within the linear operating range of the feedback loop components. We verify this by varying the levels of the stimulus and check that the frequency response of the system remains unchanged.

In this experiment, 3 levels of stimulus: 100mV, 300mV, and 500mV are applied. We obtain the same impulse response depicted in Fig. 3.2. On this basis, the system modeled is indeed operating within the linear operating range.

c. Experiment 3

The Nyquist Stability Criterion in Eqns. 1.1(a) and 1.1(b) states that oscillations will occur if the magnitude response of the OLTF is greater than unity and the phase response is a multiple of 2π . We will now use this well-established criterion to verify whether the model obtained from the system identification method can precisely predict the oscillation frequency.

In this experiment, the feedback path is fixed and we adjust the gain of the forward path to increase the magnitude response of the OLTF. We increase the linear gain of the DSP in 2 dB steps until oscillation just occurs.

In Fig. 3.6, when the gain is increased to 17 dB, oscillation is detected and measured. The power spectrum is illustrated in Fig. 3.6 with solid line. When we decrease the gain to 15 dB, the oscillation disappeared and this is also depicted. In Fig. 3.6, there are two frequency ranges where magnitude response is above 0 dB. One is from 2 kHz to 2.5 kHz while the other is from 3 kHz to 3.6 kHz. In the two frequency ranges, only the frequency around 2490 Hz meets the phase condition of the Nyquist Stability Criteria for oscillation to occur and is indicated by the measured power spectrum of the oscillation signal. From this experiment, we have shown that the model obtained by the system identification method can be used to predict oscillation frequency precisely.

Fig. 3.6 Examination of the Nyquist criterion for feedback squealing

For completeness, we will now give an example of an application of a kind of system identification method and the Nyquist Stability Criteria to predict the maximum gain in several commercial digital hearing instruments. The sequence commences with the hearing instrument generating a sweep signal (level optional) and the response is recorded. The frequency response of the feedback path is derived according to some criteria. Thereafter according to the Nyquist Stability Criteria, the maximum gain of the hearing instrument can be determined to avoid acoustical feedback. Essentially, the method to determine the frequency response of the feedback path is a special case of system identification method.

3.3 Modeling an Entire BTE Hearing Instrument System

As discussed in Chapter 1, the continuous feedback cancellation system estimates and cancels the feedback signal continuously without interrupting the normal operation of the hearing instrument. The feed-forward path response, which is determined by a set of parameters, affects the performance of feedback cancellation algorithms significantly. For this reason, it is desirable that the modeling of the entire hearing instrument system be obtained to determine the parameter set in the forward path.

In this section, we will describe the construction of a BTE digital hearing instrument and model it with the system identification method. We do this by dividing the system into five sub-blocks and system identification method is thereafter applied to each sub-block. Finally, the entire model of the BTE system comprising the cascade of the five sub-blocks is derived and the model is verified experimentally.

3.3.1 Experimental Setup

The construction of the BTE hearing instrument is illustrated in Fig. 3.7. The BTE hearing instrument includes a BTE shell, a receiver, a hook, a plastic tube, an ear mould with an adjustable vent (Select-A-Vent), and a microphone. The receiver and microphone are connected to a DSP board via conditioning circuits, similar to the connection described in Fig. 3.1. The BTE hearing instrument is mounted on the ear of a manikin and is illustrated in Fig. 3.8.

Fig. 3.7 BTE Hearing Instrument with the ear mold

Fig. 3.8 The experimental setup of the BTE digital hearing instrument system

For convenience of presentation, the sketch of the hearing instrument system is depicted in Fig. 3.9. Loudspeaker (m) serves as the sound source, and is connected to a

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PC (n). The microphone (a) in the BTE shell picks up the sound and sends it to the DSP board (e) comprising an ADC, a DAC and a DSP chip via a signal conditioning circuit (d). The DSP board (e) is connected to another PC (o), which is used to program the DSP. The processed signal by the DSP is send out to the receiver (b) via the signal conditioning circuit (d). The sound output from receiver (b) runs to the ear canal (j) through the hook (f), the plastic tube (g) and the ear mould. The sound pressure level around the real-ear coupler is picked up by a microphone (k) and transferred to the PC (n) via a conditioning amplifier (1). This is the major path where the sound transfers to the ear canal. There is still another path for sound source to reach the ear canal and this is via the vent directly (without being amplified by the hearing instrument), and this is termed the 'direct path'. Note that part of the sound output from the receiver (b) leaks out through the vent in the ear mould and is picked up by the microphone (a) again, and this path is the primary path for acoustic feedback.

Fig. 3.9 Digital BTE Hearing Instrument System

A: microphone b: receiver c: hearing aid shell d: conditioning circuit e: DSP core f: hook
g: plastic tube h: ear mould i: manikin j: ear simulator k: microphone
l: conditioning amplifier m: speaker n, o: computer

For convenience of presentation, Fig. 3.9 is redrawn in Fig. 3.10. In Fig. 3.10, a, b, di, do, e, and i represents the same as that in Fig. 3.9. S indicates the loudspeaker (m in Fig. 3.9) to generate sound as input of the hearing instrument. M indicates the ear canal microphone (k in Fig. 3.9) which serves to measure the sound pressure level in the ear canal.

Fig. 3.10 Block diagram of the hearing instrument system

S: loud speaker a: microphone d: conditional circuit b: receiver
e: DSP board h: microphone in the ear simulator i: conditioning amplifier

3.3.2 Model each Sub-block of the Entire Digital Hearing Instrument

To facilitate the modeling, we redraw the block diagram of the system as depicted in Fig. 3.11. The entire system is divided into five blocks: the input block ($I(z)$), output block ($O(z)$), feedback block ($F(z)$), direct block ($D(z)$) and DSP block ($G(z)$). The input block $I(z)$ includes the loudspeaker (S), the acoustic path from speaker to the microphone of the hearing instrument, the microphone (a) and the conditioning circuit (di). The output block includes the conditioning circuit (do) and the receiver (b), the acoustic path from receiver (b) to the microphone in the ear simulator (M), and the conditioning amplifier (i). The feedback path includes the conditioning circuit (do), the receiver (b), the acoustic feedback path, the microphone (a), and the conditioning circuit (di). The direct path includes the loudspeaker (S), the acoustic path from the speaker to microphone in the ear simulator, the microphone (M), and the conditioning amplifier (i). The DSP block includes only the algorithms embodied in the DSP core (e), including a feedback reduction algorithm.

Fig. 3.11 The hearing instrument system divided into five sub-blocks

The system identification method described earlier in Section 3.2 is applied to model each sub-block and a FIR structure is chosen for all five blocks. The impulse response of each block, except the DSP block, is depicted in Fig. 3.12. Note that the DSP block embodies a known digital processor/algorithm and does not need to be modeled.

In Fig. 3.12, the head taps which are all zeros caused by the group delay of the feedback path and the tail taps whose magnitudes are contributed by room reverberation (with almost zero value) are not plotted for clarity. The impulse response of each block is normalized separately: 0.43 for feedback block, 0.11 for input block, 7.78 for output block and 0.10 for direct block. With the four impulse responses, the whole system's model is obtained by combining them together (see later).

Fig. 3.12 Impulse response of each block

3.3.3 Modeling the Entire System

From Fig. 3.11, the complete transfer function of the system can be easily derived as:

$$H(z) = \frac{Y(z)}{X(z)} = \frac{I(z)G(z)O(z)}{1 - G(z)F(z)} + D(z) \quad - (3.5)$$

Based on the individual impulse responses of the sub-blocks, their transfer function can be easily derived. Then the transfer function $H(z)$ can be derived according to Eqn. (3.5).

3.3.4 Experimental Verification

To verify the entire hearing instrument model, two experiments are performed. In the first experiment, a fixed input signal (loudspeaker m) and the output from the conditioning amplifier are recorded simultaneously. In other words, the entire system is considered as a black box and we can achieve the impulse response of the system by the system identification method. Theoretically, our model of the complete hearing instrument system should match the results from measurements. The DSP block $G(z)$ can be modeled as any hearing loss compensation algorithms, such as a linear amplifier and compression systems. For convenience of presentation, we model $G(z)$ as a linear amplifier with gain G_0 , i.e. $G(z) = G_0 z^{-1}$.

To validate the system model given in Eqn. (3.1), we set G_0 to 0.01 and 0.1, which corresponds to two cases: the magnitude response of OLTF is far less than unity and close to unity, respectively. The model and the measurement results are depicted in Fig. 3.13. From Fig. 3.13, we note that the impulse responses from our model and experiment match very well when $G_0=0.01$ and match with a slight difference between taps 180 to 220 when $G_0=0.1$. The mismatch is probably attributed to improper measurements when the system is on the verge of oscillation. Under this condition, the filter length (300 taps) chosen for the model is insufficient to model the impulse response of the complete system.

Fig. 3.13 Impulse response of the entire hearing instrument system when DSP is programmed with a linear gain of 0.01 and 0.1

It is important for our model to predict the important performance parameters, such as maximum stable gain (MSG) and the oscillation frequency. According to our model, when G_o is set to 0.25, according to Nyquist Stability Criteria, oscillation will occur at frequency around 2200Hz.

The frequency response of the entire hearing instrument system depicted in Fig. 3.13, is calculated by the Welch method. In the physical experiment, we initially set the gain to zero and increased it gradually until the point of oscillation. At the point of oscillation, the gain of the DSP is exactly 0.25, thereby verifying that our model can precisely predict the MSG, which can be easily extended to predict added stable gain. Next, we record the output of the hearing instrument system and calculate its power spectrum which is also

depicted in Fig. 3.14. From Fig.3.14, we note that the oscillation frequency is precisely predicted by our model.

Fig. 3.14 Predict the oscillation frequency with our model

In summary, the models derived are precise and they would allow a good platform for the evaluation of our proposed algorithm in the next chapter.

3.4 Conclusions

In this chapter, we have applied the system identification method to model the feedback path of the ITE hearing instrument and have verified the model experimentally. The system identification method has been extended to model an entire BTE hearing instrument and the model has been experimentally verified. The models obtained have

been proven to be precise, hence appropriate as a platform for evaluating acoustic feedback cancellation algorithms.

Chapter 4 One-Step-Ahead Prediction Method for a Non-continuous Feedback Cancellation System

4.1 Introduction

In Chapter 1, we discussed the two well-established adaptive feedback cancellation systems: continuous system and non-continuous system. From a stability point of view, the non-continuous system is arguably more stable than continuous system - although the continuous system is almost universally used in practical hearing instruments. This is largely because in a non-continuous feedback cancellation system, when oscillation or quiet environment is detected, the normal operation of the hearing instrument is interrupted. Thereafter, the adaptive feedback canceller starts to track the changes in the feedback path. Finally, the adaptive feedback canceller is fixed and the normal operation of the hearing instrument resume.

It is well established that traditional non-continuous feedback cancellation algorithms perform well in quiet environments [50]. However, when oscillation is detected in a 'noisy' environment, such as when somebody is talking to the hearing instrument wearer, the performance of the traditional feedback cancellation algorithms may, in some cases, be affected. Note that to the feedback cancellation algorithm, the desired speech signal is also considered as 'noise' signal. Interestingly, when such systems embody the Wiener filter method for their analysis, the performance of the feedback canceller is not affected by the noise. However, in a practical feedback canceller, recursion algorithms, such as

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LMS method, are applied to reduce computation needs and these traditional recursion algorithms are often significantly degraded due to the environmental noise. In view of this, a recursion algorithm with better performance in noisy environment is highly desirable.

In this chapter, we propose the adoption of the one-step-ahead prediction method to the non-continuous feedback cancellation system. As this method can not only estimate the feedback signal, but also the noise input signal, it can better mitigate the undesired effects of the noise input signal.

Our hearing instrument system to study the acoustic feedback problem is depicted in Fig. 4.1. As discussed in Chapter 1, for the non-continuous feedback cancellation system, the feed-forward path is of no concern. Nonetheless, for completeness, we will also describe the feed-forward path. The implementation of the oscillation/quiet detector is then introduced. Finally, the Wiener filter method and the general LMS method are discussed respectively and the problem of the general LMS method is delineated followed by the proposed adoption of the one-step-ahead method for the non-continuous feedback cancellation system.

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Fig. 4.1 Multi-channel compression system with a non-continuous feedback cancellation

It was shown in Fig. 1.1 that different gains are needed to compensate the hearing loss [49, 55] at different frequency ranges and in many cases, to address the reduced dynamic range of hearing impaired, compression (Automatic Gain Control or AGC) is applied. In Fig. 4.1, the input signal, comprising the desired input and the feedback signal, is picked up by the microphone and converted to a digital signal by the ADC. The estimated feedback signal from the feedback canceller is subtracted from the entire input signal and the estimated desired input signal is obtained. The estimated signal is divided into 16 channels or frequency bands by the 16-channel filter bank. Each output of four channels is grouped and connected to an AGC. Finally the outputs of the four AGCs are combined and sent to the receiver through the DAC and amplifier. This 16-channel AGC system is a typical 16-channel hearing instrument, and in this case, emulates the Siemens Triano BTE [47].

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The estimated input signal is applied as input to the oscillation/quiet detector. Once oscillation or quite is detected, the normal operation of the hearing instrument system is interrupted by switching to the noise generator and a noise is inserted as the stimulus to the adaptive feedback canceller.

We will now present the implementation of the Oscillation/Quiet detector. Thereafter, we will present the one-step-ahead prediction method and evaluate its efficacy over the traditional method. Finally, we give a conclusion to this chapter.

4.2 Oscillation/Quiet Detector

The oscillation detector estimates the power of the largest sinusoidal component in the estimated desired signal by an adaptive notch filter [51] located in a side branch of the processing system. The decision is based on comparing the estimated power in the detected sinusoid with a threshold. The power in the detected sinusoid is calculated as the power input to the filter minus the power output from the filter, with the input and output power estimates averaged by a one-pole low-pass filters. The equations for the adaptive notch filter are listed below:

$$\begin{aligned}
 q(n) &= s(n) + \rho a(n-1)q(n-1) - \rho^2 q(n-2) \\
 r(n) &= q(n) - a(n-1)q(n-1) + q(n-2) \\
 a(n) &= a(n-1) + 2\mu q(n-1)r(n) \\
 d(n) &= \beta d(n-1) + (1-\beta)[s^2(n) - r^2(n)]
 \end{aligned}
 \tag{4.1}$$

where

$s(n)$	input to the notch filter
$r(n)$	output from the notch filter
$a(n)$	adaptive filter frequency coefficient
$q(n)$	internal sequence

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ρ	radius of the poles ($\rho=0.95$)
$d(n)$	estimated power in the oscillation frequency
μ	step size
β	parameter between 0 and 1

Note that the adaptive notch filter has only one parameter $a(n)$, which is related to the oscillation frequency. The $d(n)$ is the estimated power in the oscillation frequency and if $d(n)$ is greater than a given threshold, the system assumes that oscillation occurs and the normal operation of the system would be interrupted.

To verify the function of the adaptive notch filter, we design an experiment with Simulink, depicted in Fig. 4.2. In Fig. 4.2, the adaptive notch-filter acts not only as the oscillation detector, but also as part of the forward path if oscillation is detected. When this happens, the sound source is input to the adaptive notch filter and then to the gain block. If no oscillation is detected, the input signal is sent to the gain block directly. In the former case, the notch filter tracks and subtracts power of the oscillation frequency and sends the residual signal to the gain block.

To simulate the changes in the feedback path, one of the three feedback path models with a different magnitude response and phase response is selected. The linear gain $G = G_0 z^{-1}$ for the gain block is chosen. According to the Nyquist Stability Criteria, the oscillation frequency of the system corresponding to the three selectable feedback path models set at 2 kHz, 4 kHz, and 5.4 kHz, respectively, can be determined.

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Fig.4.2 Simulink model to validate the function of the adaptive filter

In the first instance, feedback path 1 (2 kHz) is selected and the gain G_0 is set to ensure that there is no oscillation. The gain is then increased to a higher level to ensure that oscillation occurs. The notch of the adaptive notch filter is observed to move to 2 kHz and the stable frequency response of the adaptive notch filter is illustrated in Fig. 4.3(a). We then change the feedback path model to feedback path 2. The observed frequency response of the adaptive notch filter is illustrated in Fig. 4.3(b). Finally we choose feedback path 3 and the result is illustrated in Fig. 4.3(c).

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(a)

One-Step-Ahead Prediction Method for Non-continuous Feedback Cancellation System

(b)

(c)

Fig. 4.3 Frequency response of the adaptive notch filter (a) feedback path 1, (b) feedback path 2, and (c) feedback path 3

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With this experiment, we verify (in this simulation) that the adaptive notch filter works well to track the oscillation frequency.

We will now succinctly describe the quiet/noise detector. The basic idea in the quiet detection is to calculate the average energy of the input signal and compare it to a preset threshold. If the averaged energy is less than the threshold, the quiet detector assumes the environment is quiet.

4.3 Proposed one-step-ahead prediction feedback canceller

Fig. 4.4 illustrates the basic structure of the proposed non-continuous feedback cancellation system. The meaning of the symbols in Fig. 4.4 is the same with those in Fig. 1.7. The feedback canceller, $W(z)$, adjusts its coefficients to estimate the feedback path, $F(z)$. The well established least mean square (LMS) method, which can be considered as a practical implementation of Wiener filter method, is probably the most popular feedback cancellation method used in hearing instrument systems for its robust and simplicity features. However, the LMS method suffers degradation from the input signal $x(n)$, which can be the desired speech signal or the environmental noise. To appreciate the undesired effects of the input signal $x(n)$, we will first analyze the solution with the Wiener filter method. Thereafter, we will introduce the LMS method and show how the LMS method suffers from $x(n)$. Finally, we apply the one-step-ahead method to the non-continuous feedback cancellation system to reduce the affects of $x(n)$.

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Fig. 4.4 Basic structure of feedback cancellation system

4.3.1 Wiener Filter Method

The Wiener filter method [50] can be used to analyze the performance of the feedback cancellation system and/or applied to implement the feedback cancellation system.

In the Wiener filter method, the performance (cost) function is chosen to be

$$\zeta = E[|e(n)|^2] \quad - (4.6)$$

where $E[\cdot]$ denotes statistical expectation and $e(n)$ is the error signal.

For the hearing instrument system, the input signal $u(n)$ in Fig. 4.4 is a signal with real value. For convenience of presentation, we assume that the adaptive filter $W(z)$ is a transversal filter

$$W(z) = w_0 + w_1 z^{-1} + \dots + w_{N-1} z^{-(N-1)} \quad - (4.7)$$

where $\mathbf{w} = [w_0 \ w_1 \ \dots \ w_{N-1}]^T$ is the coefficient vector.

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In this case, we can achieve the optimum coefficient vector $\mathbf{w}_o$ by setting the partial derivatives of ζ with respect to every tap weight w_i ($i=0, 1, \dots, N-1$) to zero:

$$\mathbf{w}_o = \mathbf{R}^{-1} \mathbf{p} \quad - (4.8)$$

where

$$\begin{aligned} \mathbf{R} &= E[\mathbf{u}(n)\mathbf{u}^T(n)] \\ &= \begin{Bmatrix} r_{00} & r_{01} & r_{02} & \cdots & r_{0,N-1} \\ r_{10} & r_{11} & r_{12} & \cdots & r_{1,N-1} \\ r_{20} & r_{21} & r_{22} & \cdots & r_{2,N-1} \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ r_{N-1,0} & r_{N-1,1} & r_{N-1,2} & \cdots & r_{N-1,N-1} \end{Bmatrix} \end{aligned} \quad - (4.9)$$

$$\mathbf{p} = E[\mathbf{u}(n)s(n)] \quad - (4.10)$$

where $\mathbf{u}(n) = [u(n) \ u(n-1) \ \cdots \ u(n-N+1)]^T$

It is clear that matrix $\mathbf{R}$ is unrelated to the input signal $x(n)$. As the microphone input signal $s(n)$ is the sum of the input signal $x(n)$ and the feedback signal $f(n)$, we can rewrite Eqn. (4.10) as

$$\begin{aligned} \mathbf{p} &= E[\mathbf{u}(n)s(n)] \\ &= E[\mathbf{u}(n)[x(n) + f(n)]] \\ &= E[\mathbf{u}(n)x(n)] + E[\mathbf{u}(n)f(n)] \end{aligned} \quad - (4.11)$$

As the signal $u(n)$, which is a white noise in non-continuous feedback cancellation system, is uncorrelated with the input signal $x(n)$, the first item of Eqn. (4.11) is zero. Hence, we get

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$$\mathbf{p} = E[\mathbf{u}(n)f(n)] \quad - (4.12)$$

From the above analysis, it can be seen that the solution from the Wiener filter method is not affected by input signal $x(n)$. In other words, the optimum coefficient vector $\mathbf{w}_0$ can match the corresponding part in the feedback path response $F(z)$ without the bias problem.

4.3.2 Least-Mean-Square Filter Analysis

The conventional LMS algorithm is essentially a stochastic implementation of the steepest descent algorithm. It simply replaces the cost function (Eqn. (4.6)) by its instantaneous coarse estimation:

$$\zeta = e^2(n) = (s(n) - u(n)W(z))^2 \quad - (4.13)$$

According to the concept of the steepest descent algorithm, we can achieve a recursive function to achieve an estimation of coefficient vector $\mathbf{w}$:

$$\mathbf{w}(n+1) = \mathbf{w}(n) - \mu \nabla e^2(n) \quad - (4.14)$$

where μ is the step-size parameter and ∇ is the gradient operator defined as the column vector:

$$\nabla = \begin{bmatrix} \frac{\partial}{\partial w_0} & \frac{\partial}{\partial w_1} & \dots & \frac{\partial}{\partial w_2} \end{bmatrix} \quad - (4.15)$$

Note that the i^{th} ($i = 0$ to $N-1$) element of the gradient vector $e^2(n)$ is

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$$\frac{\partial e^2(n)}{\partial w_i} = 2e(n) \frac{\partial e(n)}{\partial w_i} \quad - (4.16)$$

Also, note that $e(n) = s(n) - f'(n)$ and $s(n)$ is independent of w_i ,

$$\frac{\partial e^2(n)}{\partial w_i} = -2e(n) \frac{\partial e(n)}{\partial w_i} \quad - (4.17)$$

Substituting

$$f'(n) = \mathbf{u}(n)W(z) \quad - (4.18)$$

into Eqn. (4.17):

$$\frac{\partial e^2(n)}{\partial w_i} = 2e(n)u(n-i) \quad - (4.19)$$

Together with Eqns. (4.15) and (4.19), we obtain:

$$\nabla e^2(n) = -2e(n)\mathbf{u}(n) \quad - (4.20)$$

Substituting this result in Eqn. (4.13), we get:

$$\mathbf{w}(n+1) = \mathbf{w}(n) + 2\mu e(n)\mathbf{u}(n) \quad - (4.21)$$

This is the LMS recursion, where it suggests a simple procedure for the recursive adaptation of the filter coefficients after the arrival of every new input sample, $u(n)$, and its corresponding desired output sample, $s(n)$.

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According to the above derivation, the coefficient vector $\mathbf{w}$ can be achieved with three simple steps:

- a. Calculate $f'(n)$ according to Eqn. (4.18),
- b. Calculate error signal $e(n)$ directly by equation $e(n) = s(n) - f'(n)$, and
- c. Calculate $\mathbf{w}(n)$ according to Eqn. (4.19) for one iteration of the LMS algorithm.

To simply explain the undesired effects of the input signal $x(n)$, we re-express the error signal $e(n)$ as:

$$e(n) = s(n) - f'(n) = x(n) + f(n) - f'(n) \quad - (4.22)$$

Combining Eqns. (4.21) and (4.22), it is clear that the input signal $x(n)$ can result in an undesired bias $2\mu x(n)\mathbf{u}(n)$ to the coefficient vector $\mathbf{w}$.

It is also clear that the bias effect is determined by three parameters: step-size μ , input signal $x(n)$, and vector $\mathbf{u}(n)$. To reduce the noisy-coefficient problem, we can just simply decrease the value of step-size μ or vector $\mathbf{u}(n)$. However, this approach will increase the time needed for $\mathbf{w}$ to convergence to an optimized value, which means the normal operation of the hearing instrument will be interrupted for longer time. From a hearing instrument user's perspective, an added delay is usually unacceptable.

4.3.3 One-Step-Ahead Prediction Method

From Fig. 4.4, the input signal $s(n)$ can be expressed as:

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$$s\{n\} = f\{n\} + x(n) = F(z)u(n) + H(z)v(n) \quad - (4.23)$$

The input signal $x(n)$ is modeled as the output generated by passing a white noise $v(n)$ to an auto-regression (AR) filter $H(z)$, where $H(z)$ is a monic transfer function:

$$H(z) = \frac{1}{1 + \sum_{k=1}^N h(k)z^{-k}} \quad - (4.24)$$

The idea behind the one-step-ahead prediction method is to estimate $H\{z\}$ as well as $F(z)$. In this manner, the undesired effect of the input signal $x\{n\}$ can be reduced. According to the one-step-ahead prediction method [23], the optimal predictor of $s(n)$ equals

$$s'(n) = H^{-1}(z)F(z)u(n) + [1 - H^{-1}(z)]d(n) \quad - (4.25)$$

where $d(n) = f(n) + x(n) = F(z)u(n) + H(z)v(n)$.

Using Eqn (4.25), the prediction error $e_p(n) = s(n) - s'(n)$ can be written as

$$e_p(n) = H'(z)^{-1}[d(n) - F'(z)u(n)] \quad - (4.26)$$

A standard criterion for $H'(z)$ and $F'(z)$ is to minimize the mean square error (MSE):

$$E[e_p^2(n)] \quad - (4.27)$$

For convenience of presentation, we replace $H^{-1}(z)F(z)$ by $B(z)$ and $H^{-1}(z)$ by $A(z)$.

As $H(z)$ is monic, $A(z) = H^{-1}(z)$ is also monic:

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$$A(z) = 1 + z^{-1} \bar{A}(z) \quad - (4.28)$$

Substituting Eqn (4.28) into Eqn (4.25), we get:

$$s'(n) = B(z)u(n) + z^{-1} \bar{A}(z)(-d(n)) \quad - (4.29)$$

The feedback cancellation system based on one-step-ahead prediction method is depicted in Fig. 4.5. According to LMS criterion, $B(z)$ and $A(z)$ can be calculated. $H(z)$ and $F(z)$ can be derived from their relationship to $B(z)$ and $A(z)$.

Fig. 4.5 Feedback cancellation system based on one-step-ahead prediction method

4.3.4 Performance Parameter

To compare the performance of the one-step-ahead prediction method to the traditional LMS method, a performance parameter is defined as:

$$\sigma = \sqrt{\sum_{i=1}^r (\mathbf{f}_i - \bar{\mathbf{f}})^T (\mathbf{f}_i - \bar{\mathbf{f}}) / (r-1)} \quad - (4.30)$$

where $\bar{\mathbf{w}}$ is the averaged adaptive filter coefficient vector, which can be considered as an optimized coefficient vector, and $\mathbf{w}_i$ is the coefficient vector of the i^{th} sample process,

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and r is the number we use to calculate the average coefficient vector. A large σ indicates the system is severely affected by bias problem.

4.3.5 Auto Regression model of the speech signal

As mentioned before in this chapter, to non-continuous feedback cancellation system, the desired speech signal is a special case of noise signal. Speech-shaped noise is applied in the simulation as the input signal $x(n)$. Most audio signals, especially speech signals, can be closely approximated by auto regression (AR) model [48]. The AR model used here is a 12th order model whose frequency response is depicted in Fig. 4.6.

Fig. 4.6 Spectrum of the AR model

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4.3.6 Simulation Results

In this section, we will first show the noisy-coefficient problem caused by the external noise in the traditional LMS method by simulations. Thereafter, we will compare the performance of the traditional LMS method and the proposed one-step-ahead method.

The time domain model (impulse response) of the feedback path (indicated as Im-FB) used in the simulation is illustrated in Fig. 4.7, which is obtained by the system identification method described earlier in Chapter 3. The one sample instantiated impulse response (indicated by w) achieved by the normal LMS method is also overlaid in Fig. 4.7 for comparison. It is clear that the sample instantiated impulse response is a poor match to the feedback path model due to the noise introduced by the input signal $x(n)$. The averaged impulse response, however, can match the feedback path model considerably better, which is indicated by $w\text{-ave}$ in Fig. 4.7.

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Fig. 4.7 Impulse response estimated with normal LMS

This simulation shows that it is misleading to compare the performance of feedback cancellation algorithms with the averaged impulse response as the actual feedback canceller is unable to match the feedback path model compared to the averaged impulse response which matches well.

Fig. 4.8 depicts the impulse response of the feedback path (indicated by Im-FB), averaged impulse response by traditional LMS method (indicated by w-ave), and two root variance curves which are calculated by adding and subtracting σ from the averaged impulse response (indicated by w+ and w-), respectively. For comparison, we also illustrate the results from the one-step-ahead prediction method in Fig. 4.9.

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Fig. 4.8 Mean and root variance of impulse response from traditional methods

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Fig. 4.9 Mean and root variance from one-step-ahead method

From Figs. 4.8 and 4.9, it is clear that the one-step-ahead prediction method can reduce the bias problem significantly. Put simply, the proposed one-step-ahead prediction method is considerably less affected by the input signal

4.4 Conclusions

We have evaluated the traditional LMS feedback cancellation system and have shown that it suffers from the noisy-coefficient problem. By means of the application of one-step-ahead prediction method, we have shown that noisy-coefficient problem is considerably reduced by estimating the input signal $x(n)$ as well as the feedback signal $f(n)$. The efficiency of our method has been demonstrated by a simulation.

Chapter 5 Conclusions and Future Work

5.1 Conclusions

The primary objective of the research program is to model hearing instruments and to propose an algorithm to mitigate the acoustical feedback problem in hearing instruments, thereby allowing an increase in the gain of the hearing instrument before the onset of acoustic feedback.

We have constructed an analogue BTE hearing instrument and investigated the contribution of each component in the hearing instrument to the acoustic feedback problem by modeling each component in the feedback path and thereafter combining them to obtain the entire model of the OLTF. The model of each component in the OLTF was determined either experimentally, by theoretical analysis or from data sheets. The model of the entire OLTF has been verified against physical measurements. This two-port method has been found to be cumbersome for modeling hearing instruments and have limited application to digital hearing instruments.

A digital ITE hearing instrument has been constructed and the model of its feedback path has been developed for simulating our proposed one-step-ahead prediction feedback cancellation algorithm. The system identification method was employed to derive the model and the model was verified experimentally.

Although the feedback path model of digital hearing instrument is sufficient for simulating our proposed non-continuous feedback cancellation algorithm, we have extended the system identification to model an entire digital BTE hearing instrument and

verified its model experimentally for future work. The models obtained have been proven to be precise.

Finally we have evaluated the noisy-coefficient problem in the traditional LMS method and have applied the one-step-ahead prediction method to largely resolve this problem. By means of the application of the one-step-ahead prediction method in a non-continuous feedback cancellation algorithm, we have shown that noisy-coefficient problem is considerably reduced. The efficiency of our method has been demonstrated by simulation results.

5.2 Suggested Future Work

The suggested future work based on our current research is listed below:

- a. As mentioned in Chapter 1, the continuous feedback cancellation system is almost universally used in practical hearing instruments as the normal operation of the hearing instrument is uninterrupted. From the modeling of the entire digital hearing instrument system for simulating continuous feedback cancellation algorithm, we suggest the development of a continuous feedback cancellation system that do not suffer from the bias problem and is robust in noisy environments.
- b. The simulation of the feedback cancellation algorithm proposed is only the first step in the solution for the acoustical feedback problem. For a practical application, where it is realized in an actual hearing instrument, we suggest the following:

- i) Design a wearable digital hearing instrument,
 - ii) Implement the multi-channel compression system in the wearable digital hearing instrument,
 - iii) Test the multi-channel compression system in an anechoic room,
 - iv) Implement the one-step-ahead non-continuous feedback cancellation system in the wearable digital hearing instrument,
 - v) Test the feedback cancellation system in anechoic room, and determine its efficiency in terms of the added stable gain before the onset of feedback.
 - vi) Test the feedback cancellation in practical environments, such as in a noisy canteen.
- c. Despite the more prevalent application of continuous feedback cancellation algorithms, non-continuous feedback cancellation algorithms may yet be promising due to its potential of a lower bias problem. We suggest a more thorough investigation into the merits and disadvantages of both methods, and a possibility of combining them - in a hybrid form, perhaps.

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