Fundamental Concepts in EMG Signal Acquisition

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Preface

The field of electromyography research has enjoyed a rapid increase in popularity in the past number of years. The progressive understanding of the human body, a heightened awareness for exploring the benefits of interdisciplinary studies, the advancement of sensor technology, and the exponential increase in computational abilities of computers are all factors contributing to the expansion of EMG research. With so much information and so many different research goals, it is often easy to overlook the intricacy, the exactitude and the finesse involved in recording quality EMG signals.

This paper presents fundamental concepts pertaining to analog-to-digital data acquisition, with the specific goal of recording quality EMG signals. The concepts are presented in an intuitive fashion, with illustrative examples. Mathematical and theoretical derivations are kept to a minimum; it is presumed that the reader has limited exposure to signal processing notions and concepts. For more aggressive descriptions and derivations of these ideas, the reader is directed to the suggested Reading List and References found at the end of the document.

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1. Introduction: What is “Digital Sampling”?

Virtually all contemporary analyses and applications of the surface electromyographic signal (SEMG) are accomplished with algorithms implemented on computers. The nature of these algorithms and of computers necessitates that the signals be expressed as numerical sequences. The process by which the detected signals are converted into these numerical sequences “understood” by computers is called *analog-to-digital conversion*. Analog signals are voltage signals that are *analogous* to the physical signal they represent. The amplitude of these signals typically varies continuously throughout their range. The analog-to-digital conversion process generates a sequence of numbers, each number representing the amplitude of the analog signal at a specific point in time. The resulting number sequence is called a *digital signal*, and the analog signal is said to be *sampled*. The process is depicted in *Figure 1*, with a sample Motor Unit Action Potential (MUAP) obtained with a DE-2.1 electrode.

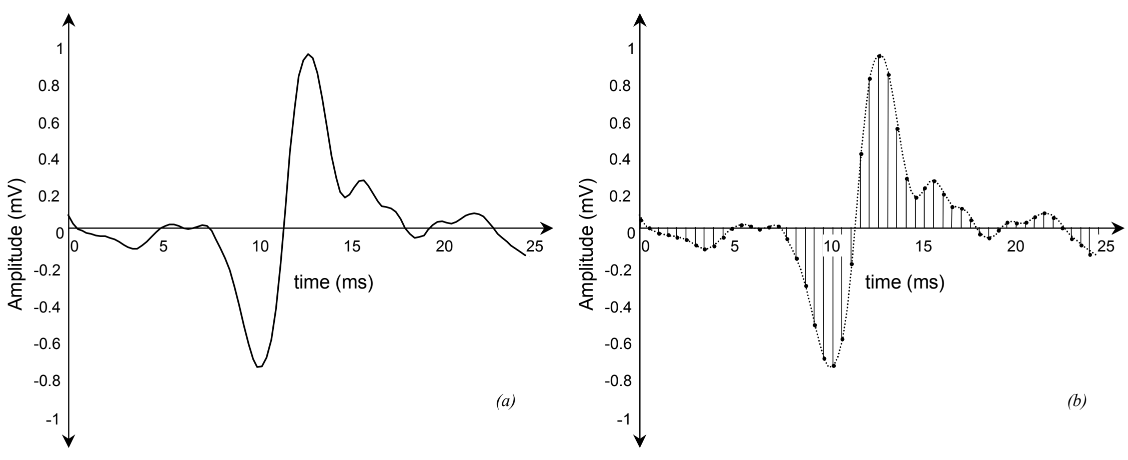


Figure 1: a) A typical analog EMG signal detected by the DE-2.1 electrode. (b) The digital sequence resulting from sampling the signal in (a), at 2 kHz (every 0.5 ms).

1.1 The Sampling Frequency

The process of signal digitization is defined by the concept of the sampling frequency. *Figure 1 (b)* depicts the sampling of an analog signal at a regular time interval of 0.5 ms. An alternate way of expressing this information is to say that the signal is sampled at a frequency of 2000 samples/second. This value is obtained by taking the inverse of the time interval, and is typically expressed in Hertz (Hz). The sampling frequency then, is said to be 2kHz. This parameter plays a critical role in establishing the accuracy and the reproducibility of the sampled signal.

1.2 How High Should the Sampling Frequency Be?

It is critical to know what the minimum acceptable sampling frequency of a signal should be in order to correctly reproduce the original analog information. The mathematical derivation of the answer to this question can be found in most introductory signal processing text books (refer to Bibliography for suggestions). We will approach the answer to this question from an intuitive perspective by considering the sampling of a simple sinusoid, as shown below in *Figure 2 (a)*

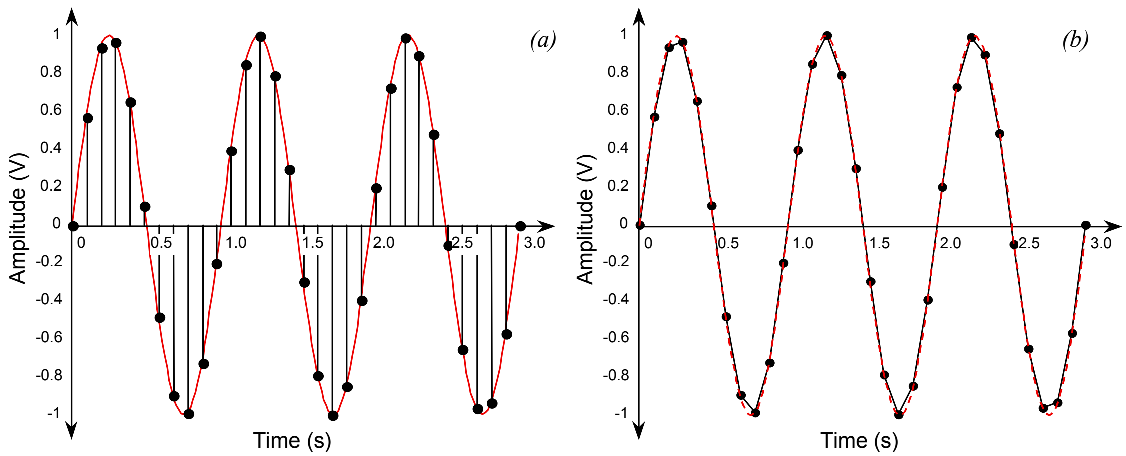


Figure 2: (a) Sampling a 1 V, 1 Hz sinusoid at 10 Hz. (b) Recreating the sinusoid sampled at 10 Hz.

This particular sinusoid can be described by its 1-Volt amplitude and its frequency of 1 Hertz. Sampling this signal at frequency of 10 Hz yields a sequence of data points that closely resemble the original sinusoid when they are connected by a line (*Figure 2*). It is fundamental to note that the lowest-frequency sinusoid capable of tracing all the sampled points is the original 1 V, 1 Hz wave.

Now consider *Figure 3(a)* where the same 1 V, 1 Hz sinusoid is sampled at a much lower frequency of approximately 2 Hz. Connecting the resulting group of data points with lines does not recreate the visual image of the original sinusoid (*Figure 3(b)*). However, if it assumed that the sequence of points must be matched with a sinusoid of the lowest possible frequency, then the only possible sine wave described by these data is the original 1 V, 1 Hz signal. The original information has been retained in the sampled sequence of points.

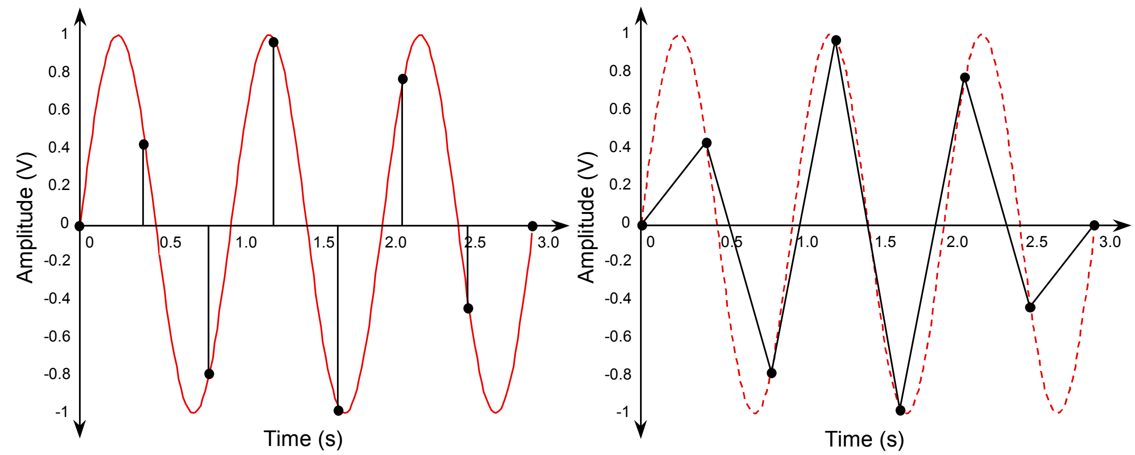


Figure 3: (a) Sampling a 1 V, 1 Hz sinusoid at approximately 2 Hz. (b) Recreating the sinusoid sampled at ~2 Hz.

1.3 Undersampling- When the Sampling Frequency is Too Low

Consider one final time the same 1 V, 1 Hz sinusoid, this time sampled every 0.75 seconds (4⁄3 Hz). Unlike the previous two cases, the resulting lowest frequency sinusoid that passes through this sequence of points is not a 1 Hz sinusoid, but rather a 1⁄3 Hz sine wave. It is clear from this example that the original signal is *undersampled* as not enough datum points have been gathered to capture all the information correctly. This condition of undersampling is said to result in *aliasing*

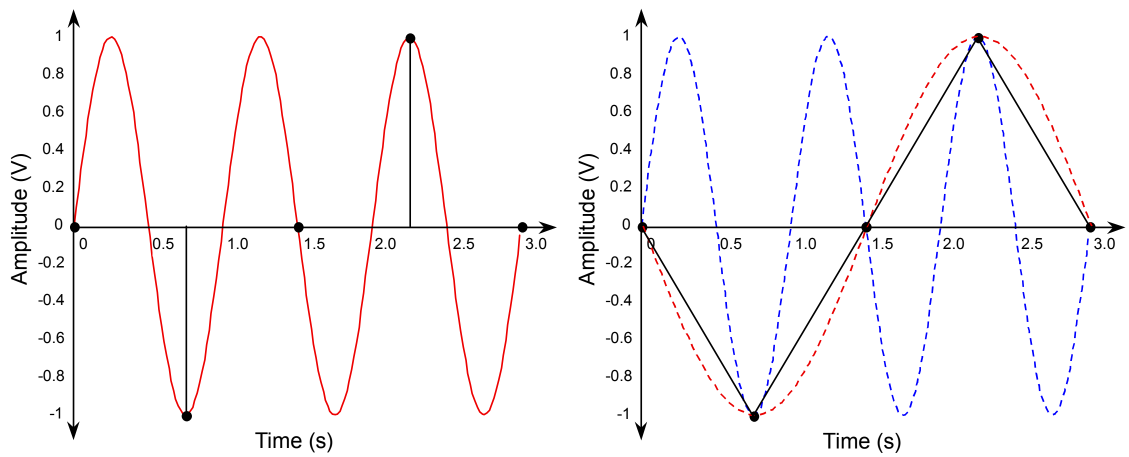


Figure 4: (a) Sampling a 1 V, 1 Hz sinusoid at 4⁄3 Hz. (b) Recreating the sinusoid sampled at 4⁄3 Hz yields the red signal at 1⁄3 Hz. The original 1 Hz signal is undersampled.

1.4 The Nyquist Frequency

With the aid of the illustrations above, it is important to realize that a sinusoid can only be correctly recreated if it is sampled ***at no less than twice its frequency***. This rule is known as the *Nyquist Theorem*. Violating the Nyquist Theorem leads to an incorrect reconstruction of the signal, typically referred to as *aliasing,* which is described later. Although these examples have been illustrated with simple sinusoids, the Nyquist Theorem holds true for all complex analog signals as is shown in the following sections.

1.5 Delsys Application Note

Delsys EMG equipment is designed to aid the user in performing high quality signal recordings and acquisitions. EMGworks software offers distinct advantages when used in conjunction with Delsys recording instrumentation. Delsys' user interface will always suggest the optimal sampling frequency to be used for a given experimental setup so that the Nyquist Theorem is always fulfilled. When using EMGworks software with third-party hardware, it is the responsibility of the user to asses the minimum sampling frequency necessary. The following chapters will provide guidelines for the user so that these assessments can be made correctly and efficiently.

2. Sinusoids and the Fourier Transform

The above example is illustrated with sinusoidal signals because they carry special significance in the science of signal processing. It can be shown that any real continuous signal can be expressed as an infinite sum of weighted sinusoids. This set of sinusoids is called a Fourier Series, the derivation and properties of which are far beyond the scope of this article. The trigonometric expression for this series is given below:

$$x(t) = A + \sum_{n=1}^{\infty} [B_n \cos(f_n \cdot t) + C_n \sin(f_n \cdot t)] \qquad \text{(Eq. 1)}$$

$$x(t) = A + B_1 \cos(f_1 \cdot t) + C_1 \sin(f_1 \cdot t) + B_2 \cos(f_2 \cdot t) + C_2 \sin(f_2 \cdot t)\ldots \qquad \text{(Eq. 2)}$$

The co-efficient labeled "*A*" represents any DC component that the signal may have (i.e. non-zero mean), the B_n and C_n represent unique coefficients for the amplitude of each cosine and sine term, while the f_n represents the unique frequency of each cosine and sine term.

2.1 Decomposing Signals into Sinusoids

Let's consider a sample analog signal, similar in appearance to a surface recorded MUAP, shown in the red trace at the bottom of *Figure 5*. This trace can be decomposed into a series of sinusoids derived from the Fourier Series described above. The first 10 sinusoids from the resulting series are shown in *Figure 5*. The summation of these 10 sinusoids is depicted in the blue trace at the bottom of the figure. It is clear from comparing this trace to the original red one that a faithful recreation of the signal can be made with only 10 sinusoids. Naturally, the fidelity of the recreated signal increases as higher frequency sinusoids are included.

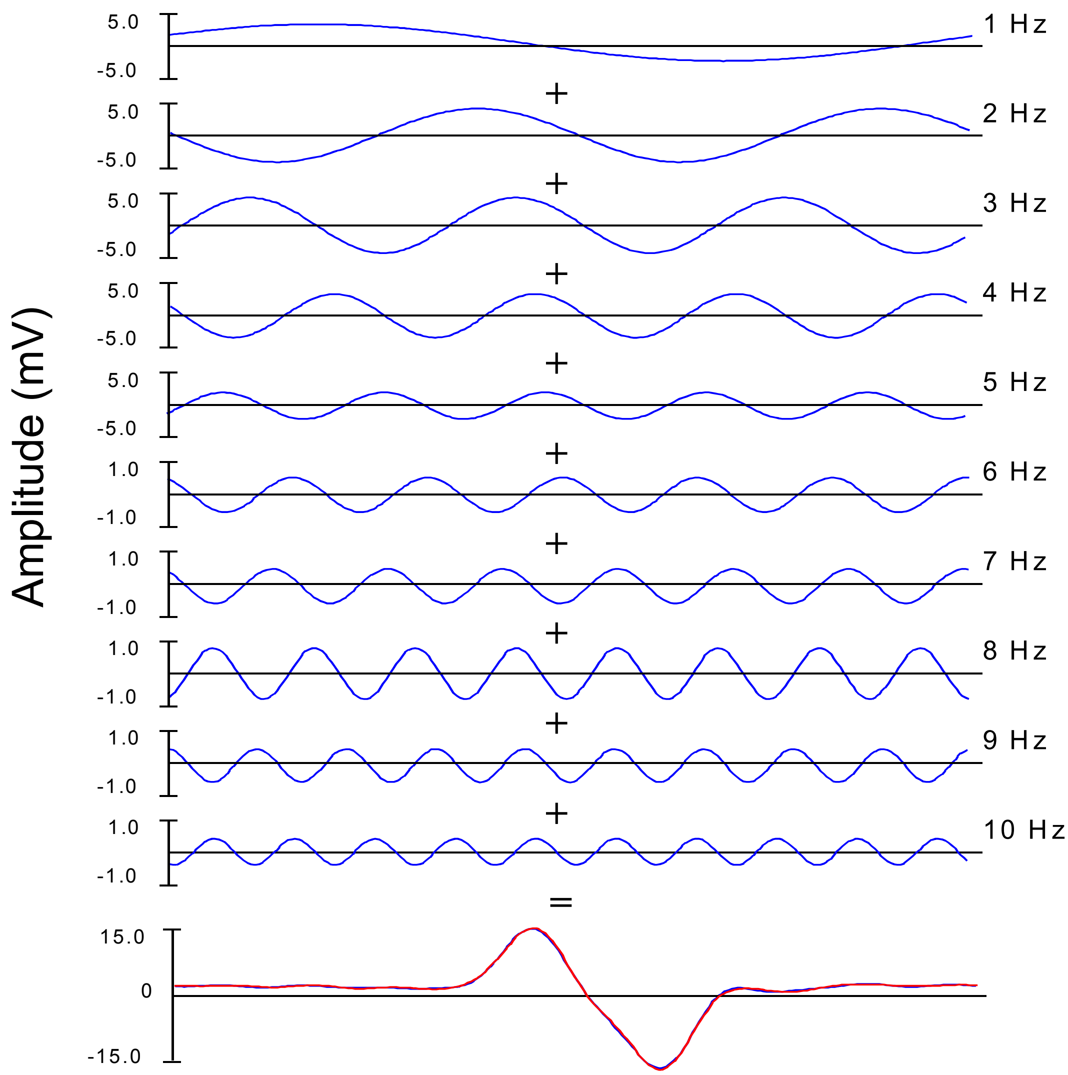


Figure 5: Fourier decomposition of a sample motor unit action potential (MUAP) recorded using a DE-2.1 electrode. The original signal is shown in red. The superposed blue signal is the mathematical summation of the 10 sinusoids above. The exact reconstruction of the red signal would require an infinite number of sinusoids, but appreciable accuracy can be achieved with only 10.

2.2 The Frequency Domain

The information depicted in *Figure 5* can be alternatively expressed in a convenient fashion by plotting a histogram of the amplitudes for each sinusoid. This concept is depicted in *Figure 6*, showing the frequency of the sinusoids on the “X-axis” and their corresponding amplitudes on the

"Y-axis". In this fashion, it is possible to describe the complete set of sinusoids that compose the electrical signal.

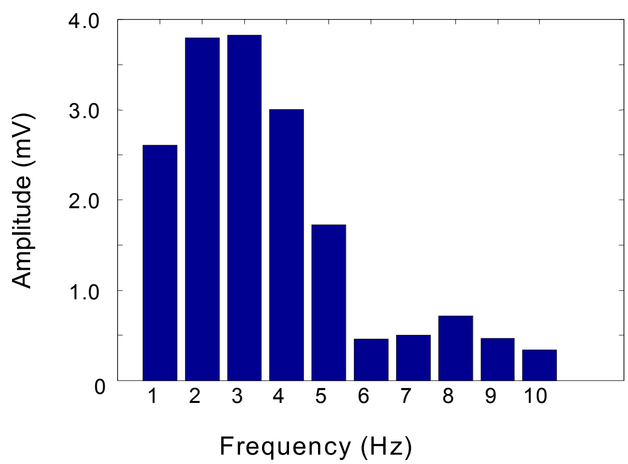


Figure 6: Amplitude histogram of the 10 sinusoids depicted in the above example of Figure 5. Note that with this method of displaying the amplitudes, it becomes readily apparent that the signal contains strong frequency components between 1 Hz and 5 Hz.

The original trace shown in *Figure 5* is said to be expressed in the "Time Domain", since it describes a voltage signal as a function of time. *Figure 6* describes the same signal in the "Frequency Domain", since it describes the amplitude of the frequencies contained in it. This type of graph is commonly called a Frequency Spectrum or Power Spectrum. Numerous algorithms and techniques have been devised over the years for extracting frequency information from time varying signals. The most basic and most popular algorithm for accomplishing this task is the "Fast Fourier Transform" or "FFT". It is highly recommended to fully understand the conditions, assumptions and caveats of the FFT and similar algorithms before employing them in data analysis.

2.3 Aliasing- How to Avoid It

The above discussion illustrated that in order to capture correctly a sinusoidal signal, the sampling rate needs to be at least twice that of the signal's frequency. Since it was also stated that all continuous analog signals can be expressed as a summation of sinusoids, it follows that the sampling frequency for any signal should be at least twice the value of the highest frequency component in the signal. Referring back to the example described in *Figure 6*, the minimum acceptable sampling frequency to capture all the relevant information of this signal is 20 Hz, since the highest frequency component is 10 Hz.

Consider the same signal described in *Figure 6* with the addition of an unwanted noise component at 13 Hz. This is shown in *Figure 7* by the red frequency component.

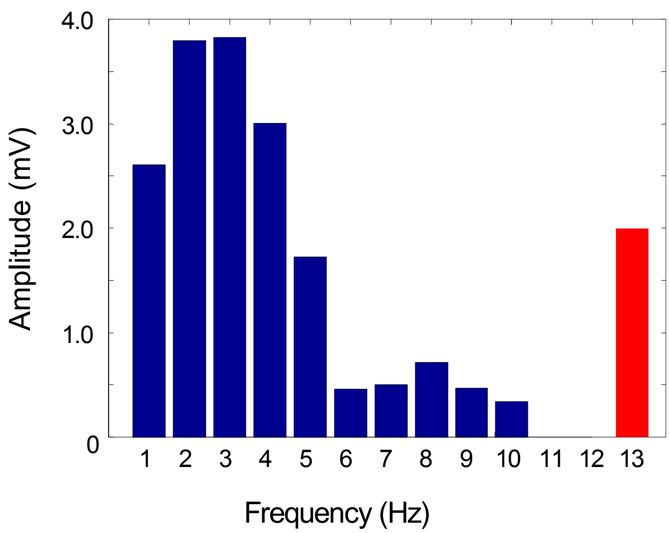


Figure 7: The frequency spectrum of the sample signal described in Figure 6 with the superposition of an external frequency not present in the original signal, shown in red. This frequency component appears at 13 Hz and is considered to be noise.

In order to digitize correctly all the information in this signal, it is necessary to sample it at a frequency of at least 26 Hz. If this is not done, and the signal is undersampled at 20 Hz, the information is captured incorrectly as shown in *Figure 8*. Note that frequencies below ½ the sampling rate (i.e. 10 Hz) have been correctly captured, but the 13 Hz component is aliased, appearing as a component "folded back" at a frequency of 7 Hz, and changing the original amplitude of this component.

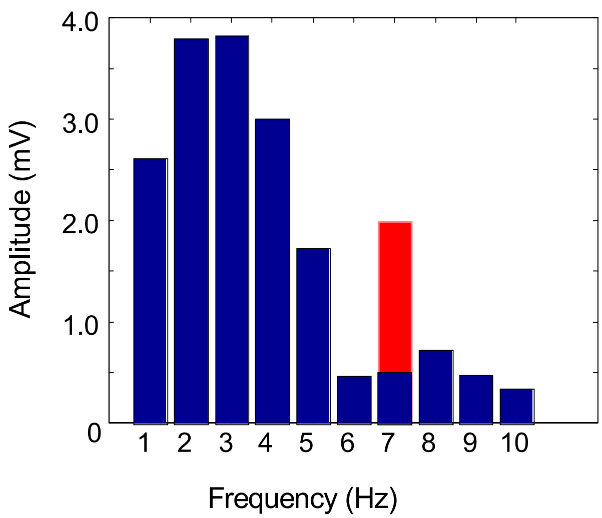


Figure 8: The aliasing of the 13 Hz noise component (shown above) due to undersampling of the signal at a frequency of 20 Hz. Note the noise appears at the incorrect frequency of 7 Hz. This component is said to be aliased

2.4 The Anti-Aliasing Filter

In order to avoid the undesirable effect of aliasing, an anti-aliasing filter is employed before the signal is sampled. It is necessary to know the bandwidth of the signal of interest in order perform this task. For example, an anti-aliasing filter with a bandwidth of 10 Hz can be applied to the signal in the above example, effectively removing the 13 Hz noise component. Once this is accomplished, the signal can be sampled at 20 Hz with no negative consequences.

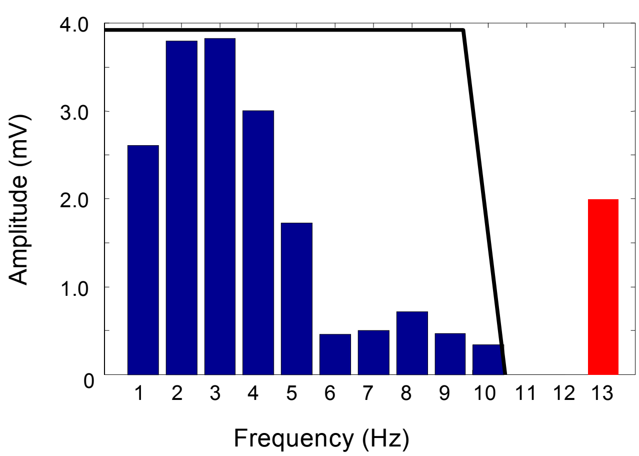


Figure 9: Removal of the 13 Hz noise component using a sharp filter with a cutoff at 10 Hz. A filter used for discarding any frequencies not in the sampling range of interest is called an antialiasing filter.

The alternative way to correctly sample all the information depicted in *Figure 9* is to set the anti-aliasing filter to a cutoff after 13 Hz, and sample the signal at a frequency of 26 Hz. This captures all of the information present, including the noise, which could be removed with a digital filter at a later point in time.

The use of an anti-aliasing filter is of paramount importance when sampling any signal. The effects of aliasing cannot be undone, nor can their presence always be detected. In any A/D acquisition system, the cutoff frequency of the antialiasing filter must always be less than one half the sampling frequency. This guarantees that the no aliasing will occur.

2.5 Delsys Practical Note

The full bandwidth of the Surface EMG signal spans up to 500 Hz. All standard Delsys equipment is designed and configured to optimally detect the complete spectrum of the EMG signal. All systems have built-in anti-aliasing filters, with upper bandwidths of 500 Hz. In typical circumstances, the detected SEMG signal will contain little energy above 400 Hz, however it is strongly recommended that sampling of the EMG signal is performed at least at 1000 Hz, as dictated by the Nyquist Theorem. Sampling the EMG system outputs at a rate less than 1000 samples/second may irreparably distort the signal due to aliasing. The default minimum sampling rate of EMGworks Signal Acquisition and Analysis Software is 1024 Hz. Extreme care must be exercised to preserve signal integrity when modifying any hardware or software default parameters.

3. Filters

The antialiasing filter has been introduced as the specific case of a low pass filter, an essential component of any digitization process. In practice, there is often need for other types of filters, some involved with proper signal conditioning (implying the use of "analog filters") others necessary for the analysis of data once it has been digitized ("digital filters"). The study of filter theory and its applications is a science in its own right, the subject of which is appropriately left to the countless texts describing it. The following discussion is a presentation of some basic filter concepts commonly encountered (and often misunderstood) in the study of Electromyography. Once again the reader is urged to review the references for more sophisticated discussions on the subject.

3.1 The Ideal Filter Types

A filter is a device designed to attenuate specific ranges of frequencies, while allowing others to pass, and in so doing limit in some fashion the frequency spectrum of a signal. The frequency range(s) which is attenuated is called the *stopband*, and the range which is transmitted is called the *passband*. The behavior of filters can be characterized by one of four functions depicted in *Figure 10*: low-pass, high-pass, band-bass and band-stop.

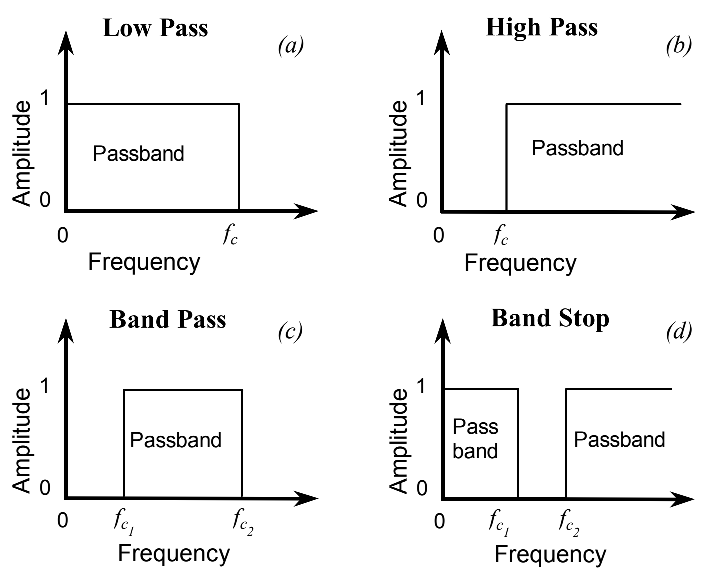


Figure 10: The four basic filter types. Frequencies where the filter response amplitude is 1 are defined as passband regions, while frequencies where the filter response amplitude is 0 are defined as stopband regions. The cutoff frequency is denoted by 'f_c'. (a) Low-Pass filter- all frequencies higher than f_c are attenuated to zero. (b) High-Pass filter- all frequencies below f_c are attenuated to zero. (c) Band-Pass filter- all frequencies lower than f_{c1} and higher than f_{c2} are attenuated to zero. (d) Band-Stop filter- all frequencies higher than f_{c1} and lower than f_{c2} are attenuated to zero.

The depictions of *Figure 10* are representations of ideal filter characteristics, typically referred to as brick-wall responses, that imply the following behavior:

1. The passband amplitude response is continuously flat at a value of 1. The frequencies which are allowed to pass through the filter do so completely undistorted.

2. The stopband amplitude response is continuously flat at a value of 0. The undesirable frequencies are completely suppressed.
3. The transition between the passband and the stopband happens instantaneously.

3.2 Ideal Phase Response

The brick-wall responses of the ideal filters of *Figure 10* only describe changes in the magnitude of the signals as a function of frequency. Since sinusoids are entirely described by the magnitude of their amplitude and by the angle of their phase, the complete specification of a filter must include a statement of its phase response. All causal filters introduce a time delay at the output since they cannot act instantaneously on the input signal. An ideal filter's time delay is independent of frequency. That is to say, the filter will modify the phase of each frequency component entering the system exactly the same way. This behavior is characterized by the phase plot illustrated in *Figure 11*

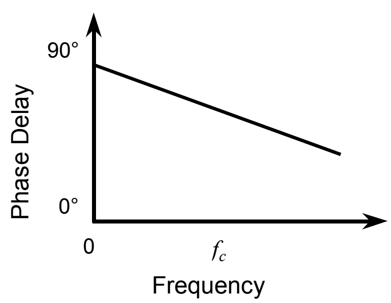


Figure 11: Figure Phase plot of an ideal filter. Note that all frequencies are affected by the same delay so that no phase distortion occurs

Practical implementations of filters in either analog or digital format can only approach the behavior described by these ideal characteristics. Furthermore, filters are usually designed to maximize the ideality of one of these characteristics, at the expense of some other. In order to gain a better appreciation for these trade-offs it is necessary to describe the filter response in a more realistic light.

3.3 The Practical Filter

The ideal filter models presented in *Figure 10* can be modified to include parameters which more accurately describe the behavior of realizable filters (whether these are analog or digital in nature). The relative flatness of the passband(s) and stopband(s) can be described by the addition of a ripple factor which specifies the maximum and minimum deviation from the ideal value. Furthermore, the transition characteristics of a filter cannot change in a discontinuous fashion between the passband and stopband as described by the brick-wall response. As such, a transition zone is defined indicating a contained region of the transmission bandwidth where the signal transmission shifts from passband to stopband or vice versa. Naturally, the narrower this transition zone is, the more ideal the filter is. The lack of a sharp "corner" in the transition zone necessitates the definition of the "corner frequency", visually defined in the brickwall response filter. Historically, this filter specification has been defined as the frequency where the power output of the filter is one-half of the input power.

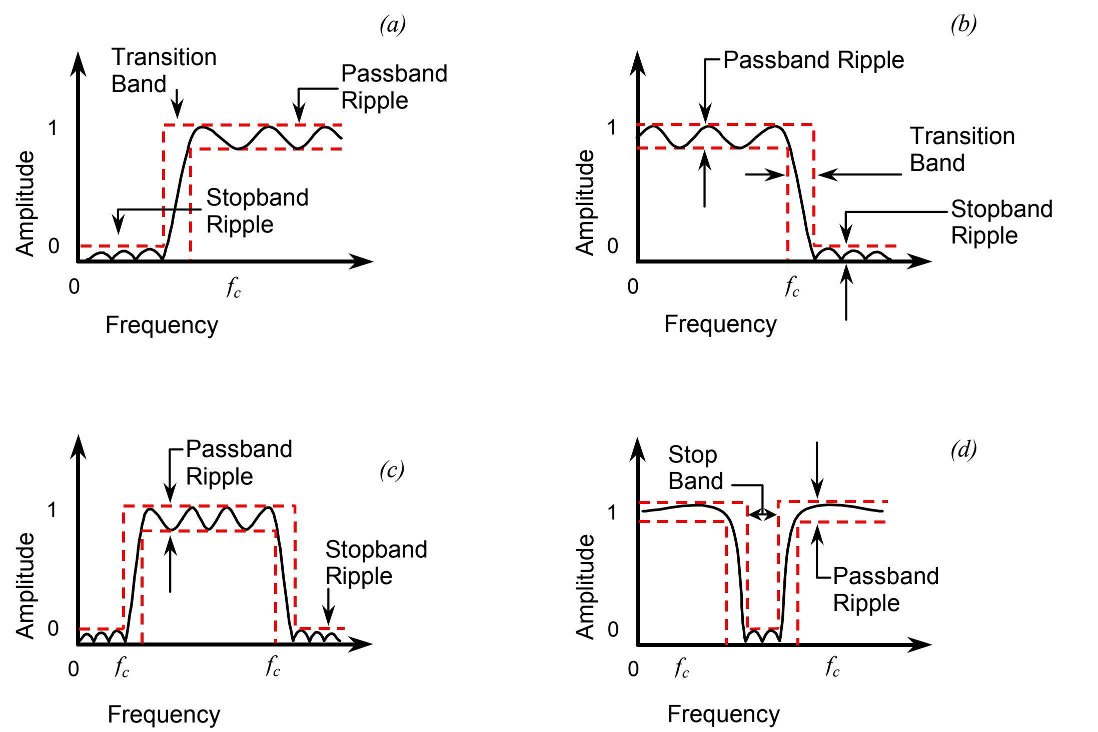


Figure 12: Characteristics of practical filters: (a) Low Pass, (b) High Pass, (c) Band Pass and (d) Stop Band. Note the specification of an amplitude ripple for all passband and stopband regions, as well as the definition of a transition zone with a specified corner frequency (f_c) and sometimes the slope of the transition band.

The relative steepness of the filter's transition zone is described by the filter's "order"; higher order filters yield narrow transition zones at the expense of complexity. Note that a filter's stop band region will never completely eliminate frequency components in this range. It is important to specify minimum acceptable attenuation factors for the specific application of the filter. Factors ranging from –20 to –100 dB (1/10 to 1/100000) are achievable.

3.3.1 Non-linear Phase Response

Many practical filter implementations suffer from a non-linear phase response. In these cases, the phase delay of the filter changes as a function of the input frequency. That is to say that different frequencies get delayed by different amounts, causing phase distortion within the signal. For the surface EMG signal, this is usually not a concern as the nature of the detection process does not permit phase preservation. Recall that the surface EMG signal is a superposition of many action potentials. This superposition renders phase information of each action potential indistinguishable from its many neighbors. Furthermore, minute changes in muscle fiber orientation, motor unit firing rates and electrode contact position may cause significant changes in phase characteristics, leading to inconsistencies between recordings, and even within the same recording. Due to the relative ineffectiveness of harnessing phase information in the EMG signal, the phase characteristics of filters typically used in Electromyography are not considered in detail when compared to the amplitude response characteristics.

3.4 Measuring Amplitude- Voltage, Power and Decibels

The filter illustrations presented in *Figure 10* and *Figure 12* show approximate amplitudes of 1 for filter transmission bands and 0 for filter stop bands. These amplitudes represent the magnitude of the filter transfer function (i.e. the gain). One parameter used for characterizing filters is the gain, expressed as the ratio of output voltage to input voltage, keeping in mind that the voltages may be expressed as time-varying functions:

$$Gain(v)_{filter} = v_{out} / v_{in} \quad \text{(Eq. 3)}$$

In the illustrations above, the gain of the filter transfer function is 1 when the filter's voltage output is the same as the input, and obviously 0 when the filter voltage output is 0.

A common way of expressing a filter's gain characteristic is with the logarithmic units of decibels (dB). A filter's voltage gain calculated in decibels is determined as follows:

$$Gain(v)_{filter} = 20log(v_{out} / v_{in})\ dB \quad \text{(Eq. 4)}$$

Note that if the ratio of filter's output voltage amplitude to it's input amplitude is 1 (i.e. gain=1), then the filter's gain in decibels is 0 dB:

$$Gain(v)_{filter} = 20log(1)\ dB \quad \text{(Eq. 5)}$$

$$Gain(v)_{filter} = 0\ dB \quad \text{(Eq. 6)}$$

Consider, for example, a filter with a gain of 100 in the passband region. Its gain in decibels for this frequency band would be:

$$Gain(v)_{filter} = 20log(100)\ dB \quad \text{(Eq. 7)}$$

$$Gain(v)_{filter} = 40\ dB \quad \text{(Eq. 8)}$$

In comparison, consider the same filter with an attenuation of 1/100 in the stop band region. Its gain for this frequency band would be:

$$Gain(v)_{filter} = 20log(0.01)\ dB \quad \text{(Eq. 9)}$$

$$Gain(v)_{filter} = -40\ dB \quad \text{(Eq. 10)}$$

Table 1 compares example gains and attenuations expressed as ratios and in decibels.

Ratio of v_{out}/v_{in}	$20\log(v_{out}/v_{in})$ (dB)		Ratio of v_{out}/v_{in}	$20\log(v_{out}/v_{in})$ (dB)
0.000001	-120		1,000,000	120
0.00001	-100		100,000	100
0.0001	-80		10,000	80
0.001	-60		1,000	60
0.01	-40		100	40
0.1	-20		10	20
0.5	-6		2	6
0.707	-3		1.413	3
1	0		1	0

Table 1: Example gain and attenuation factors expressed in decibels. Note that decibel values less than 0 denote attenuation, while positive values describe amplification.

It can be shown that a signal's power is quadratically related to it's voltage amplitude. Thus, the filter's power transfer function can be shown to be:

$$PowerGain(v) = v_{out}^2 / v_{in}^2 \qquad \text{(Eq. 11)}$$

We will make use of this fact in the next section to define the corner frequency of the filter.

3.5 The 3-dB Frequency

The previous sections presented the "brickwall" response of ideal filter characteristics as well as the need to relax the ideal characteristics due to the limitations of practical implementations. An important specification of frequency-limiting filters is the establishment of a corner frequency which demarcates the Passband and Stopband regions. In many cases this value can be defined as the frequency where the power of the filter's signal output is ½ that of its input.

$$PowerGain(v) = v_{out}^2 / v_{in}^2 \qquad \text{(Eq. 12)}$$

$$1/2 = v_{out}^2 / v_{in}^2 \qquad \text{(Eq. 13)}$$

$$1/(\sqrt{2}) = v_{out}/v_{in} \qquad \text{(Eq. 14)}$$

The corner frequency is thus defined as that frequency where the output voltage is approximately 0.707 of the input value. In Decibels, this frequency is called the “3-dB point”, as demonstrated below:

$$Gain(v)_{f_{3dB}} = 20\log(v_{out}/v_{in})\ \text{dB} \qquad \text{(Eq. 15)}$$

$$Gain(v)_{f_{3dB}} = 20\log(1/(\sqrt{2}))\ \text{dB} \qquad \text{(Eq. 16)}$$

$$Gain(v)_{f_{3dB}} = \text{-3 dB} \qquad \text{(Eq. 17)}$$

3.6 Filter Order

An additional characteristic describing the behavior of a filter is the width of the transition zone illustrated in *Figure 12*. The tighter this transition range is, the more complex the filter design must be. The complexity of the filter (and hence the steepness of the transition slope) can be characterized by the “order” of the filter.

The simplest design is a first-order filter. This filter’s transition band attenuates the input signal -20 dB for every 10-fold change in frequency. The filter is said to attenuate at -20 dB/decade. This concept is illustrated in the first-order low pass filter shown in *Figure 13 (a)*, where a decade refers to the increase in frequency by a factor of 10. Recall that –20 dB corresponds to a reduction by a factor of 10. Thus the filter will reduce the amplitude of the input signal by $1/10^{th}$ for every decade increase in frequency. This same attenuation slope can be alternately expressed as –6 dB/octave, where an octave refers to the doubling of a frequency.

To further illustrate the point, a second order low pass filter is shown in *Figure 13 (b)*. The attenuation of this design is –40 dB/decade or –12 dB/octave. While the performance of this latter design is essentially double that of the former, it comes at the expense of higher complexity- generally, the second-order filter is composed of two first order filter stages in series. The reader is directed to the suggested reading list for comprehensive information pertaining to filter design and issues of stability.

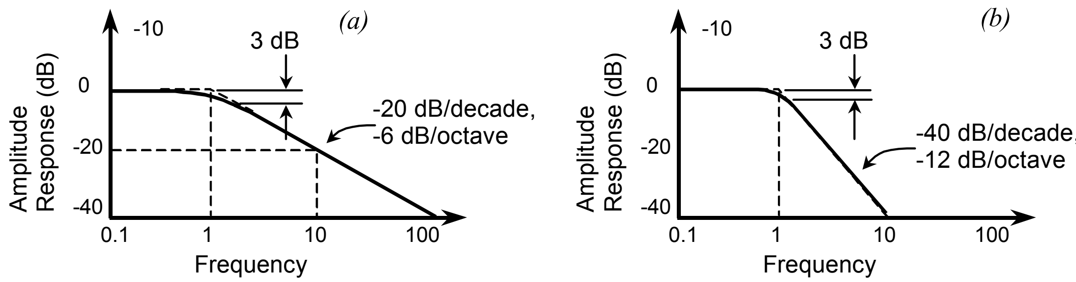


Figure 13: First-Order Low Pass Filter. Input frequencies above the corner frequency f_c are attenuated by -20 dB for every increase in frequency by a factor of 10. Recall that -20 dB corresponds to a decrease in amplitude by a factor of 10. This can equivalently be expressed as an attenuation of -6 dB/octave, where an octave refers to the doubling of a frequency. The magnitude frequency response of filters is conveniently plotted on logarithmic axis so that these attenuations can be represented by linear plots. (b) Second-Order Low Pass Filter. The slope of the magnitude response for this design shows that attenuation of the input frequencies is double that of the filter in (a).

3.7 Filter Types

Now that the basic characteristics of filters have been presented, we are in a position to appreciate the salient features of some well-establish and commonly-used filter types. Each type has specific parameters which can be optimized to approach ideal filter features at the expense of some other characteristic. Unfortunately, there is no readily available filter which can approach all the attributes of the ideal filter.

The Butterworth Filter

This filter is best used for its maximally flat response in the transmission passband, minimizing passband ripple (refer to *Figure 12*). As can be seen from the magnitude response plots of *Figure 14*, the ideal brick wall response is approached as the order “N” is increased. Specifying the maximum overshoot in the passband allows the determination of the minimum necessary order to achieve the desired response. This filter is best suited for applications requiring preservation of amplitude linearity in the passband region. It is precisely this feature which makes the Butterworth filter an ideal candidate for conditioning the EMG signal. The corner frequency, f_c, of this filter is defined as the 3-db point as described in the previous sections. Note that the phase response of this filter is not particularly linear. This filter is completely specified by the maximum passband gain, the cutoff frequency and the filter order. Refer to *Figure 14 (a)* for the magnitude and phase responses of different order Butterworth filters.

The Chebyshev Filter

Similar to the Butterworth filter, this filter can achieve steep rolloffs with high order designs. The Chebyshev outperforms the Butterworth's attenuation in the transition band for the same order design. This advantage, however, comes at the expense of noticeable ripple in the passband regions (refer to *Figure 14 (a)* and *(b)*). The total number of maxima and minima in the passband region is equal to the filter order. Unlike the Butterworth filter, the cutoff frequency for this filter is not specified at the 3-dB point, but rather at the frequency where the specified maximum passband ripple is exceeded. Like the Butterworth filter, this filter is completely specified by the maximum passband gain, the cutoff frequency and the filter order.

The Elliptic Filter

When compared to the Butterworth and the Chebyshev filters, the Elliptic filter maintains the steepest cutoff slope for the lowest filter order. The filter, however, suffers from ripple in both the passband and the stopband regions. A sharp cutoff is achieved by adding dips or "notches" in the stopband regions. These notches introduce zero transmission (i.e. complete attenuation) in selected areas. In addition to the maximum passband gain and the corner frequency, complete specification of this filter includes its order and the stopband ripple. The complexity of this filter usually necessitates the use of a computer when it is being designed. Note that the phase response of this filter is particularly non-linear. Refer to *Figure 14 (c)* for magnitude and phase response examples.

The Thompson or Bessel Filter

The magnitude response of the Bessel filter is monotonic and smooth- there are no ripples in the transmission band or the stop band. The roll off of this filter however, is much less steep than the filters presented above. The main advantage to this filter is its exceptional phase linearity. This preservation of phase also minimizes "ringing" caused by sharp inputs (known as step or impulse responses) which is commonly found in the other filters as shown in *Figure 14*.

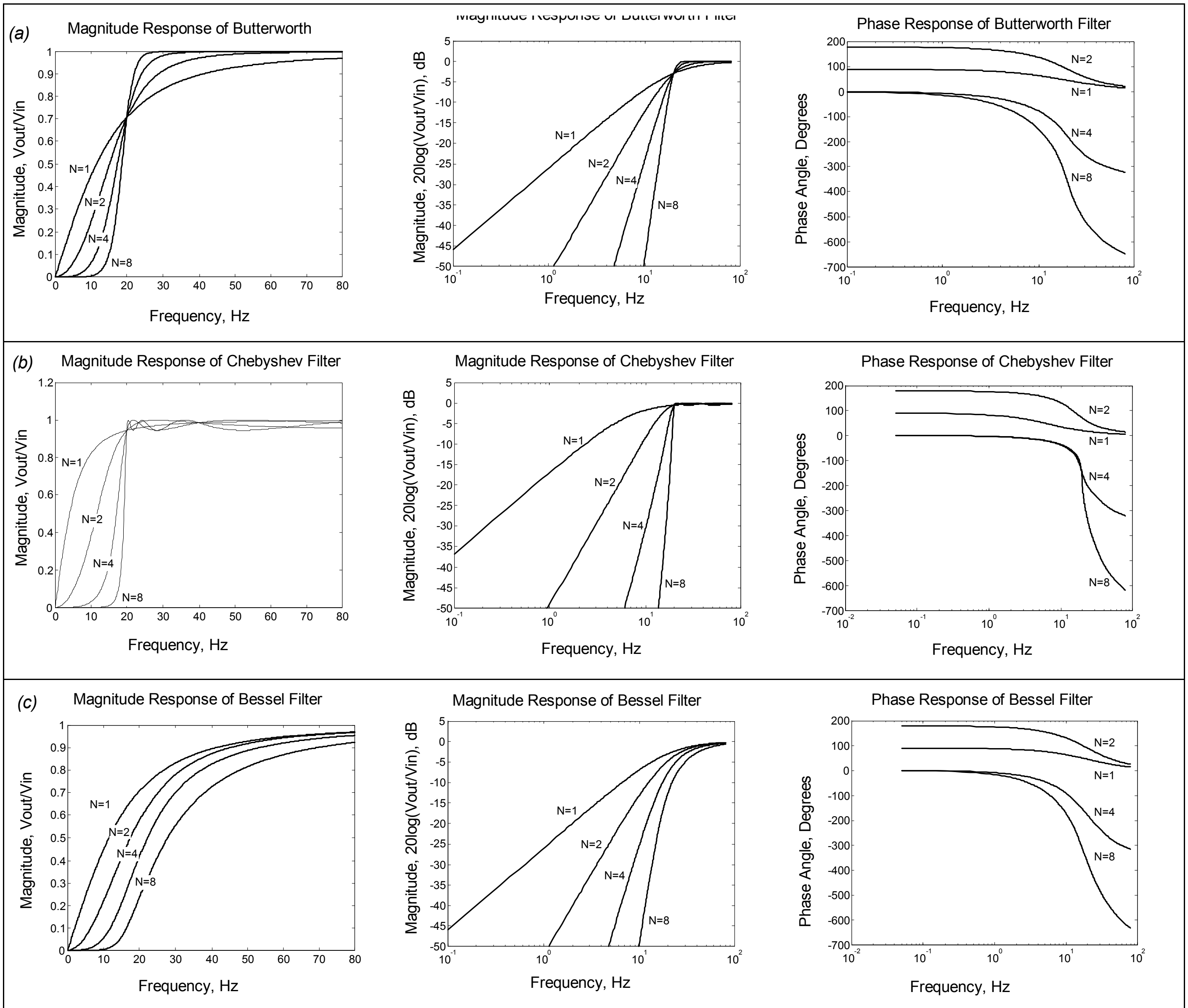


Figure 14: Magnitude and phase comparison of high-pass filter types with a cutoff of 20Hz and varying orders (n=1,2,4,8). (a) Butterworth filter, (b) Chebyshev filter, (c) Bessel Filter, (d) Elliptic Filter. Note how the logarithmic expression reveals particular behavior patterns at large attenuations. These examples demonstrate salient features of each filter type, such as linearity of pass-bands, stops and phase responses as well as relative cutoff sharpness.

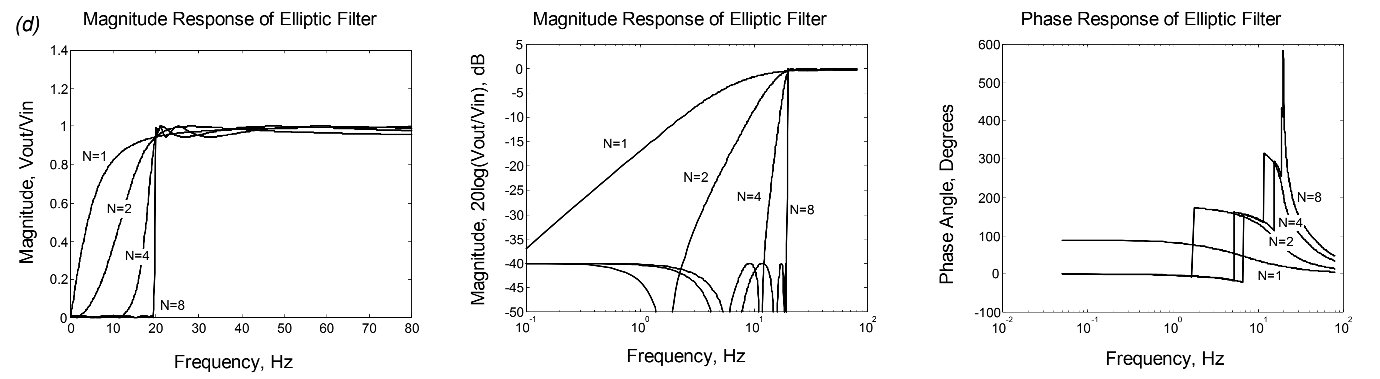


Figure 14: Magnitude and phase comparison of high-pass filter types with a cutoff of 20Hz and varying orders (n=1,2,4,8). (a) Butterworth filter, (b) Chebyshev filter, (c) Bessel Filter, (d) Elliptic Filter. Note how the logarithmic expression reveals particular behavior patterns at large attenuations. These examples demonstrate salient features of each filter type, such as linearity of pass-bands, stops and phase responses as well as relative cutoff sharpness.

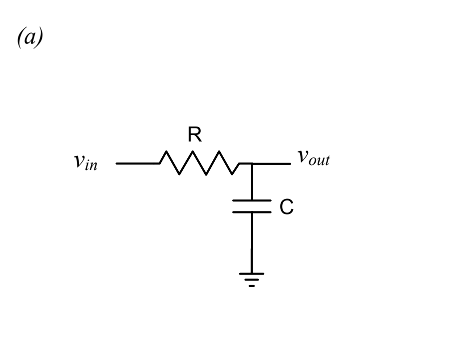
3.8 Analog vs. Digital Filters

The previous section described the behavior and the specification of several classic filter types. These filters (along with many others) can be implemented in either the analog signal domain (where the signals are continuously varying voltages) or in the digital domain (where the analog signals have been sampled and are represented by an array of numbers).

Analog Filters

Analog filters are usually implemented with electronic circuits, making use of three fundamental components: resistors, capacitors and inductors. By arranging these components in a variety of configurations, it is possible to customize filter performance to very specific needs. In addition, operational amplifiers are commonly used to increase the performance of these filters. It is important to note that these filters are commonly used in “signal conditioning stages” before any digitization takes place. Signal conditioning generally refers to the modification of a signal for the purpose of facilitating its interaction with other components, circuits or systems. This may involve the removal of unwanted noise or the reduction of bandwidth to simplify further signal analysis or processing. The most notable application of this kind is low-pass filtering for anti-aliasing purposes, as described above. It is critical for the anti-aliasing filter to be applied on the analog signal before any digitization occurs since the effects of incorrectly-sampled data cannot be undone.

The performance of analog filters is directly related to the quality of the components used and the circuit design. Things such as component tolerances, power consumption, design techniques and often the physical size components all play important roles in establishing the practical limits of analog filters.



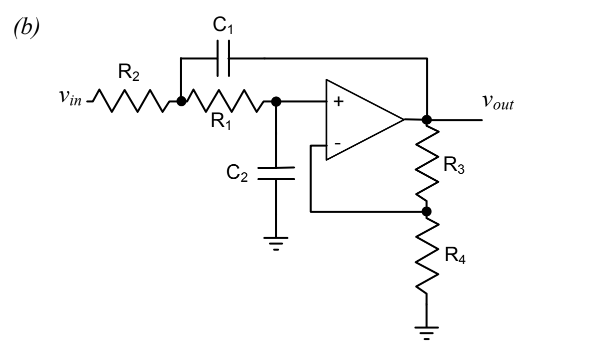


Figure 15: a) Single-pole low pass filter. This is the simplest analog filter possible with one resistor and one capacitor. The rolloffs of this basic design is 20 dB/decade, with a corner frequency given as 1/RC. b) Double-pole low pass filter. This design essentially cascades two single pole filters, facilitated by the use of an op amp. The rolloff of this filter is 40 dB/decade. The correct combination of R and C values can tailor this generic filter to have different responses, such as the Butterworth and Chebyshev filters previously described. Higher order filters are obtained by cascading more RC single pole filters.

Digital Filters

The digitization of electric signals into sequences of numbers permits the complete manipulation of these signals to occur mathematically. Voltage signals that are expressed as numbers can easily be scaled through scalar multiplication or offset by adding constants; they can be rectified by using the absolute value operator or modulated with other signals through multiplication. The digital realm provides unbounded opportunities for condition and processing of the signal. This branch of science is know as *digital signal processing.*

The digital implementation of the filters described in the previous sections is typically accomplished through various schemes of weighted averaging. An example of this process is illustrated in *Figure 16*. It begins with an analog voltage signal which is digitized with an ADC after it is appropriately conditioned. Once sampled, the signal is defined by a sequence of numbers representing the voltage amplitude at specific instances in time. A window of "*n*" input "taps" is created, each tap consecutively holding one value of the sampled data (x_n to x_i). Individual tap values are then multiplied by a specific weighting factor (h_n to h_i). The current filter output, y_i, is then calculated by summing all the weighted input tap values:

$$y_i = \sum_{k=i-n}^{i} h_k x_k \qquad (Eq.\ 18)$$

where k is the summation index, i is sample value index, n is the number filter taps, x is the filter input value, h is tap weight and y is the filter output value.

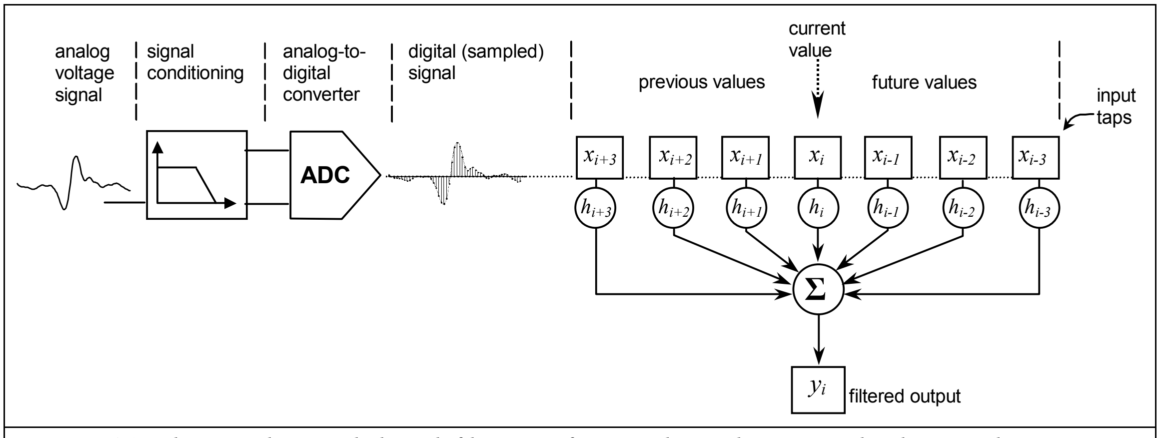


Figure 16: The sampling and digital filtering of an analog voltage signal. The signal transmission propagates form left to right starting with an analog voltage signal being first digitized and then filtered. Note that the filter output, y_i, can be a function of past values (x_{i+k}, k>0) and, if desired, of "future" values (x_{i+k}, k<0). Digital filters making use of future values are called non-causal filters.

From the illustration of *Figure 16*, an interesting property of this type of filter becomes apparent: the filter's output, y_i, can be based on the past values of x_{i+k} (k<0) as well as the future values of x_{i+k} (k>0). This possibility accounts for an added design parameter which is not available in analog filters due to their causal nature. Furthermore, digital filters can be designed to make use of previously calculated outputs. In a sense, the digital filter can be thought of as having a feedback component (y_{i-1}) used for determining the current output y_i. These filters are generally called IIR (Infinite Impulse Response) filters or Recursive Filters. In contrast, filters that do not make use of previous outputs are called FIR (Finite Impulse Response) filters. Generally, IIR filters can achieve much steeper roll-offs than FIR filters for the same number of input taps, but care must be taken to ensure the feedback in the filter does not render it unstable. FIR filters are always stable but may require many more taps than IIR filters to achieve the same results. The reader is directed to the suggested reading list for further elaborations on this topic.

3.9 Delsys Practical Note

Delsys EMG systems include high performance analog filters for signal antialiasing and instrumentation noise management. All filters are designed to maximize signal passband linearity, as well as phase linearity. Transitions bands for EMG systems are typically -80 dB/decade, with passband ripples within 0.5% deviating. Current hardware specifications can be obtained from the Delsys web site (www.delsys.com). Filters present in the EMG systems can be modified upon request, but should never be removed or excluded from the system. The analog filters present in the hardware are designed to capture the full bandwidth of the EMG signal, as well as any auxiliary signals used in conjuction with the EMG recording. Once the signals have been digitized and stored on a computer, additional filtering can be effected on the signals through EMGworks software. This filtering requires the definition of either an FIR or an IIR digital filter through the specification of tap input coefficients (weights) as derived from the desired filter's characteristics. The reader is directed to the EMGworks User's Manual for further information on this topic.

4. Considerations for Analog-to-Digital Converters

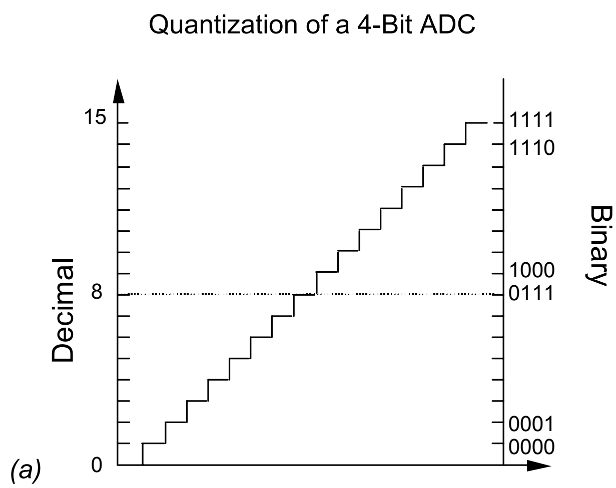
The digitization process of an analog signal is performed by a device known as an Analog-to-Digital Converter (ADC), as described in the previous sections. These devices are a common component of modern electronic products, and their use is highly varied and widespread; it is important that each application is assessed by considering the advantages and limitations of the specified ADC.

4.1 Quantization

The concept of quantization is introduced when datum values can only be represented by a limited number of digits. In the case of computers, these values are described by *binary digits*, abbreviated as "bits". All analog-to-digital converters have a fixed number of bits available for quantifying the voltage signal detected at the input. The most common Adds quantize with a resolution of 8, 12 or 16 bits, although other configurations are available. *Figure 17 (a)* illustrates the quantized range of a 4-bit ADC. Note that 4 bits can describe 16 unique values. The number of values described by and n-bit number (referred to as a 'word') is calculated with the following formula:

$$\text{Range of n-bit ADC} = 2^n \text{ values} \qquad \text{(Eq. 19)}$$

The ranges of common analog-to-digital converters are listed in *Figure 17 (b)*.



Quantization Ranges of Different-Length Binary Words

Number of Bits	Unique Values
2	4
4	16
8	256
12	4096
14	16384
16	65535

Figure 17: a) Quantization steps of 4 bit analog-to-digital converter. The steps from 0 to 15 are scaled to represent the full voltage range input range (for example, the input range could be scaled for ±1 V, where the bits from 0 to 7 represent the negative voltage range and the bits from 8 to 15 represent the positive voltage range. b) Quantization range as a function of bit length. Note that the possible resolution doubles with the addition of each bit to the word length.

4.2 Dynamic Range

The digitization of an analog voltage signal is specified over a particular range. That is to say, a maximum and minimum input voltage is defined over which the quantization should occur. By defining this range of operation for a given *n-bit* quantization scheme, the precision or "resolution" of the analog-to-digital converter can be characterized with the following equation:

$$V_{resolution} = V_{range}/(2^n) \qquad \text{(Eq. 20)}$$

For example, in the case of a 16-bit A/D system, for a range between -5 and +5 Volts, the resolution becomes:

$$\begin{aligned} V_{resolution} &= 10\text{V}/(2^{16}) \\ &= 1.53\times10^{-4}\ \text{V} \\ &= 153\ \mu \end{aligned}$$

In this case the digitization process is unable to resolve input voltage fluctuations that are less than 153 μV. This inherent limitation of a discrete number representation scheme is said to introduce "quantization error" in the measurement process. The reader is referred to the suggested reading list for a detailed analysis of quantization error. It is important to ensure that the quantization error introduced by the digitization process does not significantly affect the accuracy of the measured signal.

Typical ranges for ADCs are ±1.25, ±2.5, ±5, and ±10 Volts. As stated in the example above, a 16-bit ADC specified for a range of ±5V, would have a precision of 153 μV. This same ADC specified for a range one fourth this size, at ±1.25V would have a precision of 38 μV (a value 4 times smaller than the previous case). This point is further illustrated in *Figure 18*. It is important to ensure that the ADC range encompasses the full span of the voltage input while maintaining the minimum resolution necessary. If the necessary resolution for a given input voltage swing cannot be obtained, then it may be necessary to use an ADC with a higher number of quantization bits.

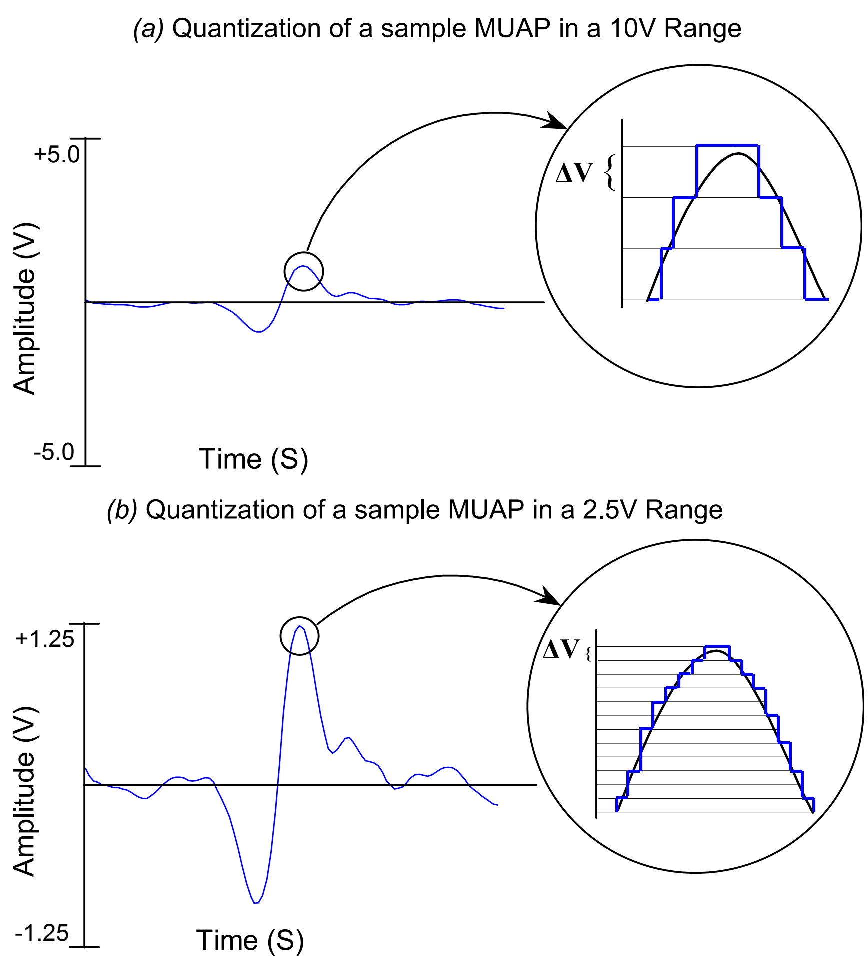


Figure 18: Resolution vs. A/D Range: (a) A representative signal with a peak-to-peak amplitude of 2.5 volts sampled over a ±5 V range. For a 16 bit digital converter, the resolution, ΔV, is 153 me. (b) the same signal sampled over a ±1.25 V range. For the same 16 bit ADC, the resolution has now increased to four-fold to 38 mV

4.3 EMG Signal Quantization

When choosing an appropriate ADC for digitizing EMG signals, it is important to consider three interacting factors:

1. The gain of the system.

2. The input noise of the system.
3. The maximum voltage output of the system.

It is important to understand how these three factors interrelate in order to determine the necessary specifications of the analog-to-digital conversion system.

System Gain

Knowledge of the overall system amplification is necessary for relating the output signals to the true input signals detected. For example, if the overall system gain is 1000 V/V, the occurrence of a 1 V spike at the output corresponds to a 1 mV (1/1000 V) disturbance at the input. A similar 1 V output recorded with a gain of 10 000 would be caused by a 100 μV (1/10 000 V) input signal. When the amplitude of output signals are divided by the overall system gain to obtain the input amplitude, it said that they are *referred to input*, abbreviated as "r.t.i.". This is a useful procedure for modeling and comparing signal and noise characteristics, regardless of a signal's origin.

System Noise

Noise can be described as any portion or aspect of the output signal which is undesirable and may possibly mask the true signal of interest. Noise can have many sources and interpretations; it can come from external radiated sources (such as "line interference"), it can be caused by electrical disturbances intrinsic to the recording environment (such as motion or stimulus artifact), and it can be caused by the nature of the recording devices themselves. The design of the EMG equipment, the establishment of a noise-free recording environment and the methodologies for using the EMG equipment must be carefully considered so that the EMG signal may be recorded with high fidelity and so that the signal-to-noise ratio is maximized.

In most cases, the degree to which external radiated noise sources and environmental noise sources affect the detected signal, can largely be controlled by the user and the recording methodology. The intrinsic noise of the EMG equipment, however, is beyond the control of the user and is grossly determined by the design and construction of the equipment. The electronic components and the design of DelSys equipment is regularly updated to include recent developments in low noise technologies. At this point in time DelSys has succeeded in obtaining an extremely low system noise of 5 μV(r.t.i.) per channel, measured by connecting the EMG electrode inputs to the reference potential. This means that if the output of a channel is recorded with no EMG signal, a baseline noise with an average amplitude of 5 μV(r.t.i.) will be observed. This is an important specification, as the minimum discernible EMG signal is within this range. A more comprehensive discussion on noise is presented in the following section on Noise.

Signal Range

The signal range of a system is defined as the maximum voltage output the device is capable of sustaining. The voltage output of Delsys EMG systems is specified to be within the range of ±5 V. This means that even in the most adverse circumstances, when the amplifiers are saturated or in the presence of excessively large noise artifacts, the system output will never exceed ±5 V.

4.4 Determining the ADC Specifications

With the information presented above, it is possible to begin assessing the ADC characteristics. The following points should be considered, not necessarily in the order presented:

ADC Range Setting

It is logical to state that the ADC range setting should match the analog output voltage of the EMG system of ±5 V. An input range larger than this 10-volt swing would offer no benefit, as the system is physically incapable of outputting signals that would take advantage of the added span. A smaller range than the rated output would increase overall signal precision, but must be used with caution due to the possibility of ADC input saturation.

Gain Setting

The selection of gain for a signal conditioning system is determined by the amplitude range of the input signals and the desired system output signal range. In the case of the EMG signal, the input amplitude ranges from ±20 μV for the faintest signals to ±2 mV for robust signals. For the purpose of interacting with common electronic instrumentation, it is convenient to amplify this EMG signal so that it never exceeds ±5 V. An amplification factor of 1000 would comfortably place the EMG signal in an amplitude range of ±20 mV to ±2 V.

Different gain settings can be used if it is known that the recorded signal of interest will occupy a more stringent subset of the presumed range. For example, if it is not expected to record EMG signals greater than ±200 μV for a particular muscle site, it may be advantageous to use a gain of 10 000, since this will place the output at ±0.2 V to ±2 V. Similarly if the detected signal is expected to be abnormally large, ±20 mV for example, a gain of 100 could be used. Choosing an appropriate gain setting so that the output signal range is compatible with the next stage of electronic equipment (usually a data acquisition system) is key for high signal fidelity.

Another essential consideration when selecting a system gain is the effect of non-linearities on the signal output, occurring most often when the input signal is outside the ideal or the expected range. For example, if an amplification factor of 10000 is used, and the input signal to the system is 1 mV, the expected output should be 10 V. However, if the system is only capable of outputting a signal in the 5-volt range, the output signal will be clipped. The recorded signal would then not be a correct linear representation of the input signal.

Other non-linearities could be introduced from the components used for electrical safety isolation. All isolation sub-systems) have specified linear operating ranges. If the gain is such that a portion of the output signal does not fall within this specified linear range, then the potential exists for the signal to be distorted (most often unpredictably). Fortunately, with the proper gain settings and distribution throughout the system, it is possible to ensure that the signal of interest is always trashiness in a linear fashion.

DelSys EMG systems have selectable gains of 100,.1000 and 10000. The recommended gain setting for most recording scenarios is 1000 V/V. As will be described in the next section, use of a 16-bit A/D system captures the full span of the EMG signal. It also guarantees operation in the linear range of the isolation transformers included with each system.

Minimum Resolution:

An important specification to determine is the minimum acceptable signal resolution. With a baseline noise of ±5 μV, it is necessary to digitize the signal with enough bits so that even the faintest EMG activity can be appreciably quantified.

A 16-bit analog to digital system set at ±5 V has a resolution of 0.153 μV(r.t.i.) for a system with a gain of 1000. This means that the recorded *noise* can be resolved with at least 5 bits (i.e. 32 quantization steps). This leaves ample resolution for all EMG activity, which will be decidedly larger than the baseline level. In contrast, a 12 bit A-D system with the same parameters has a much poorer resolution 2.441 mV. This value can be improved to 244 μV(r.t.i.) if a system gain of 10000 is used. Clearly, a 16-bit digitization system with a fixed gain of 1000 is the preferred choice for full scale range and minimum signal resolution.

Range	**Number of Bits**								
	8			**12**			**16**		
	G=100	G=1k	G=10k	G=100	G=1k	G=10k	G=100	G=1k	G=10k
±10 V	781 μV	78.1 μV	7.81 μV	48.8 μV	4.88 μV	488 nV	3.05 μV	305 nV	30.5 nV
±5 V	391 μV	39.1 μV	3.91 μV	24.4 μV	2.44 μV	244 nV	1.53 μV	153 nV	15.3 nV
±2.5 V	195 μV	19.5 μV	1.95 μV	12.2 μV	1.22 μV	122 nV	763 nV	76.3 nV	7.63 nV
±1.25 V	97.7 μV	9.77 μV	977 nV	6.10 μV	610 nV	61.0 nV	381 nV	38.1 nV	3.81 nV

Table 2: Achievable resolution as a function of available bits, ADC range setting and system gain. All calculations are referred to input (r.t.i.). Note that DelSys Systems have a baseline noise of 5 μV (r.t.i.). The recommended configuration for DelSys Systems is to digitize data at a ±5V range with a 16-bit analog-to-digital converter and a gain setting of 1000. This will ensure appropriate resolution under any condition. (Metric prefixes: 'k' refers to a factor of 10^3 (e.g. 10 k=10000), 'μ' refers to a factor of 10^{-6} (e.g. 391 μV = 0.000391 V), and 'n' refers to a factor of 10^{-9} (e.g. 153 nV=0.000000153 V)

4.5 Delsys Practical Note

All Delsys EMG equipment is typically supplied with 16-bit data acquisition capability. This policy ensures ample minimum signal resolution as well as plenty of dynamic range for a ±5V signal span. This configuration is found to be the most versatile, as it is ideal for guaranteeing the accuracy of very low voltage EMG signals while permitting the inclusion high voltage signals from auxiliary inputs such as goniometers and force gauges. Data acquisition systems with less than 16 bits may present limitations when analyzing very low level EMG contractions, while the increase in resolution of ADCs greater than 16 bits is superfluous and comes at a much greater expense.