TRANSMITTING BIOLOGICAL WAVEFORMS USING A CELLULAR PHONE

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There exists a need to remotely monitor fully mobile patients in their natural environments. Monitoring a patient's biological waveforms can track a patient's vital signs or facilitate the diagnosis of a disease, which could then be treated to help prolong and/or improve the subject's life. If a patient must be monitored without the delay associated with delivering data stored on a recording device, biotelemetry is necessary. Biotelemetry entails transmitting biological waveforms to a remote site for recording, processing and analysis. Due to the limitations of the currently popular methods of biotelemetry, this thesis proposes the use of the increasingly prevalent cellular phone system. An adaptor design is developed to facilitate biotelemetry utilizing the most common features of a cell phone, barring the need for cell phone modification, as required for affordability. As cell phones notoriously confound sensitive medical equipment, especially patient-connected devices, their use is often distanced from sensitive equipment. However, the desire to use cell phones to transmit biological waveforms requires their joint-proximity to patient-connected devices. The adaptor must amplify the waveforms while rejecting cell phone interference to achieve an adequate signal-to-noise ratio. As the frequency range of most biological data does not conform to the passband of the phone system, the adapter must modulate the biological data. To limit the adapter's size and weight, this design exploits the cell phone's battery power. Methods are also introduced to receive and reconstruct high-fidelity representations of the original biological waveform.

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1.0 BIOLOGICAL WAVEFORMS

Biological waveforms are time-varying biopotentials, representing electrical activity corresponding to bodily functions, which can be measured on the body's surface. We can monitor a patient by acquiring and transmitting desired biological waveforms. Electrocardiograms (ECG) and electroencephalograms (EEG) are the two most commonly monitored biological waveforms. As this thesis will exemplify the use of cell phones in transmitting biological waveforms using ECG and EEG, brief introductions of each follow.

1.1 ECG

The rhythmic behavior of the heart can be monitored and used as a diagnostic tool to detect heart abnormalities by acquiring the electrical activity on the body surface, across the heart, known as ECG. This electrical activity originates from the electrical activation of muscles in the heart, inherently indicating the approach of mechanical motion. The electric potentials generated by the heart appear throughout the body and can be measured across its surface. Figure 1 displays a typical ECG waveform. An ECG signal is characterized by five peaks and valleys labeled in with the successive letters P, Q, R, S, T, and U [1].

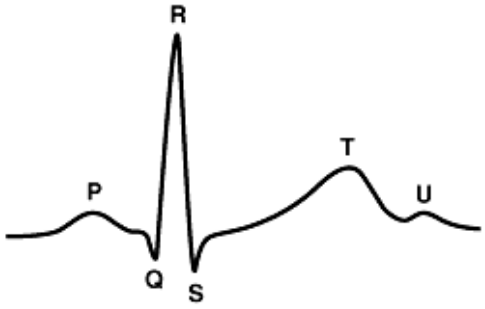


Figure 1: Typical ECG detailing the characteristic peaks and valleys

1.2 EEG

EEG are biological waveforms which can be acquired on the human scalp; these waveforms correspond to brain activity. As EEG appears random in nature there are neither characteristic peaks nor valleys. Monitoring an epileptic patient's EEG may detect the onset of seizures. Monitoring a Parkinson patient's EEG may instruct the use of medical treatment.

2.0 BIOTELEMETRY

A patient experiencing difficult-to-diagnose symptoms (suggesting cardiac arrhythmia) may be asked by a doctor to wear an ECG monitor. A currently popular method for obtaining a remote patient's biological data, while performing normal daily activities, is performed by attaching a portable recording device to the patient such as a Holter ECG monitor. The data is recorded on a memory device which can then be delivered to clinicians for review. However, if retrieving the data or the delay associated with mailing or hand delivering the data is unacceptable, biotelemetry is necessary.

Biotelemetry provides a wireless link between the subject and the remote site where the recording, signal processing, and displaying functions are performed [2]. Rather than using a traditional radio transceiver, which can only broadcast over a limited range, we suggest using the readily available cell phone to transmit biological data by creating a link between the subject and a computer receiving the signal via a landline phone [3]. This thesis discusses the methodology of transmitting biological data using a cell phone leading to the development of a cell phone adaptor to facilitate this communication link.

This technique is exemplified by first experimentally transmitting a single channel of ECG in figure 2. The ECG signal is first acquired from the subject. The cell phone adapter allows us to interface the ECG with the cell phone. The cell phone transmits to the cell phone tower, which

can then wirelessly link the subject to a remote site where the transmitted signal can be recorded and processed to recover the biological data to be analyzed.

In section 16, this single channel biotelemetry system is modeled, expanded for multi-channel systems, and simulated.

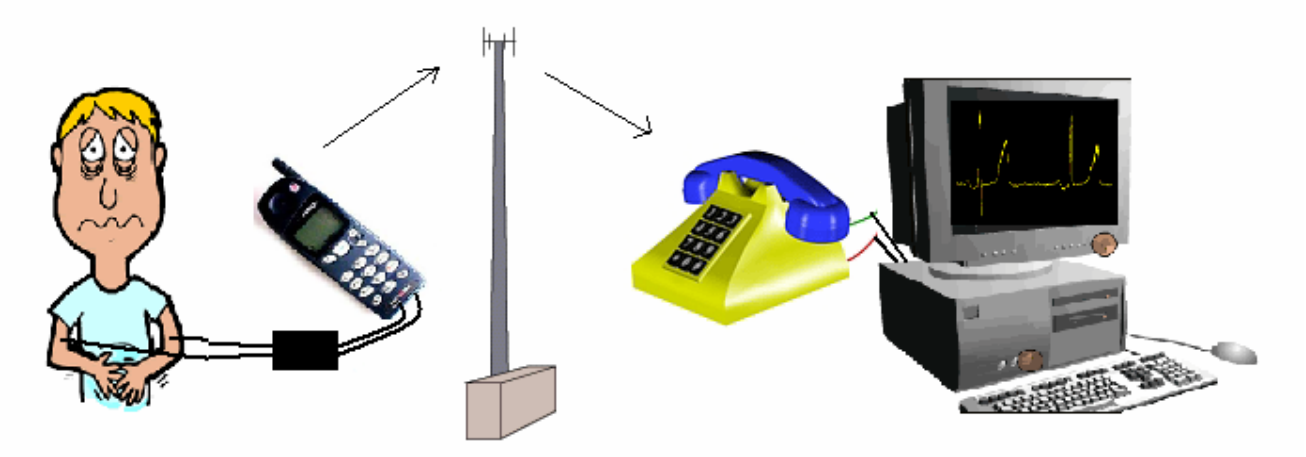


Figure 2: Using a cell phone for biotelemetry

3.0 OBSTACLES

Biological waveforms are low amplitude signals; ECG is measured around 1 mVp-p. The frequency range of biological waveforms usually lies below 200 Hz (the frequency band of ECG signal is mainly between 0.1-200 Hz) [2] which does not exist within the telephone system's passband (400-3400 Hz) [4]. While in transmission, the cellular phone creates a very noisy environment for detecting biopotentials. These high frequencies in conjunction with non-linearities in the amplifier can "demodulate" the interference to a frequency within our desired biological waveform band [5].

Normally we could simply reduce the source of interference, but in order to transmit the signal this is not possible. If the source of the interference is not reducible, it would usually be kept away from the sensitive device by law (ban on cell phones in patient areas) or by life-saving advice (keeping cell phone at least 18cm from pacemaker). However, this is not possible in our application as we need to interface our signal into an unmodified common cell phone. Patient-connected devices lessen the effectiveness of the usual means of interference control [6]. Hence we must make painstaking efforts to prevent intrusion of this debilitating interference by creatively combining multiple interference-reduction techniques. Since this device is intended for portable use it must consume low power, take up a small space, be light-weight, and be able to function without a direct earth-ground connection.

4.0 SYSTEM OVERVIEW

Shown in figure 3, is a flow chart of the simplified system outlining our systematic solution to transmitting a single channel biological waveform using a cell phone. An amplifier is needed to acquire a weak biological signal (embedded within large interference) and increase the magnitude of the signal so that it can be further processed [2]. Since biological waveforms are usually buried below 200 Hz, but the passband of the phone system is 400-3400 Hz, we need to modulate the waveform to a frequency within the phone's passband. The modulated data is interfaced to the cell phone (via the microphone input pins) to transmit the modulated signal which is received by a landline phone. A sound card acquires the signal received by the landline phone. The acquired signal is demodulated using MATLAB to release the waveform previously embedded around the carrier frequency. A representation of the original waveform may then be processed, displayed, analyzed, and saved on the receiving computer.

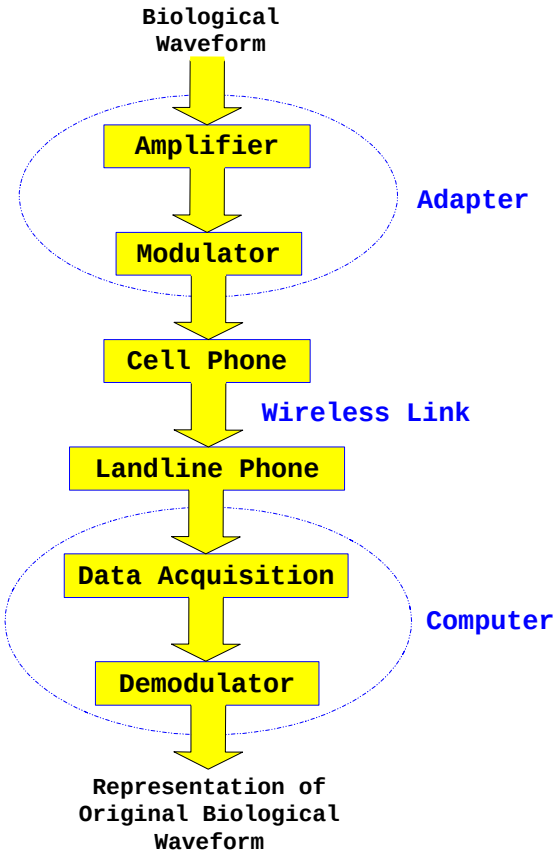


Figure 3: Simplified system flow chart

Figure 4 introduces a more detailed block diagram of the adapter designed to interface patient's biological data with the cell phone for biotelemetry. This figure is shown below to introduce the adapter system as a conglomerate of subsystems required to amplify the signal while reducing interference, modulate and filter the biological data (to facilitate transmission via the wireless link), while conditioning the cell phone's power supply to be used by the adapter (eliminating the need for a separate power supply, reducing the size and weight of the adapter). But in order to understand the components of this system and their interconnections we must begin by examining the characteristics of our desired signal, the interference inherent to patient-connected devices and cell phones, a method of interfacing most cell phones without modification, and a

method of using the cell phone to power the adapter. We will begin by characterizing our exemplary biological data.

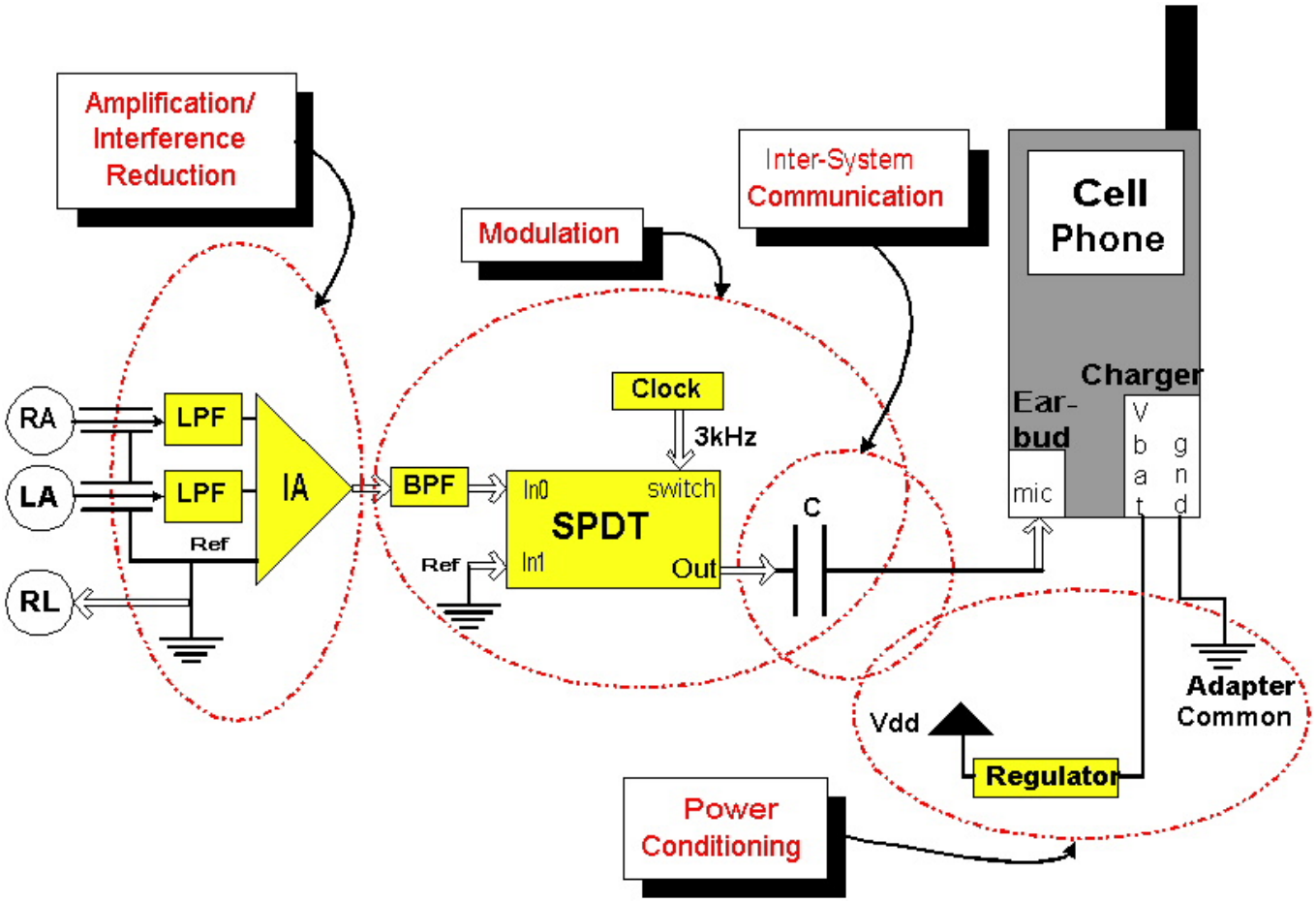


Figure 4: Cell phone adapter block diagram

5.0 BIOPOTENTIAL AMPLIFIERS

Amplifiers that are specifically designed to increase the signal strength of biological waveforms are called biopotential amplifiers. Biopotential amplifiers are required to increase the amplitude of the waveforms to facilitate signal processing, displaying, saving, or transmitting the waveforms. The type of biological waveform and the system which accepts the amplified waveform as its input dictates the required bandwidth and gain (amplification). For our application it is necessary to amplify our minute biopotentials to the required amplitude at the data interface (microphone input) of the cell phone, while rejecting unwanted signals to achieve an acceptable signal-to-noise ratio (SNR). Considering that signals common to both electrodes do not carry useful biological data, it is necessary to use a differential amplifier to amplify the voltage difference between the two electrodes while rejecting all other signals common to these electrodes (deemed interference resulting from nearby power sources or cell phones). The amplitude of our common-mode signal (interference) could reach 1V while the amplitude of the differential signal (ECG) lies around 1mV. The amplifier must therefore provide a high common mode rejection ratio (CMRR) in conjunction with a large input impedance (Z_{in}) to form a high effective CMRR (CMRRe). The CMRR is a measure of the ability of the amplifier to amplify differential signals while rejecting common mode signals ($CMRR=A_d/A_{cm}$) [2], [7], [8].

6.0 CELLULAR PHONE TECHNOLOGY

The main attraction to the cellular phone's use for biotelemetry stems from its widespread existence, low cost, improving service, and their ability to transmit (and receive) data at unlimited distance, while consuming little power. As long as the user is within the operating range of a base station, the cellular network can create an unlimited-distance wireless link between the patient and the receiving site of data acquisition, signal processing and display. This wireless link can provide signal transmission while consuming little power, which furnishes a longer lifespan for smaller batteries. This is achieved by the cellular grid system. By installing numerous communication towers to be used for relaying information to other towers or cell phones, each phone only needs to be within range of a communication tower to establish a wireless link to another cell phone. Hence, the denser the tower population the shorter the maximum distance a cell phone needs to transmit to a tower and therefore less power is needed to be emitted [4].

In the analog world, the maximum number of wireless phones which could simultaneously operate in a region was limited by the FCC-allotted communication frequency band and the bandwidth of each channel. In addition this number was cut in half, by using one channel to send signals and another channel to receive signals simultaneously for a duplex (two-way) communication channel. There are currently numerous strategies which provide ways to increase the maximum number of wireless phones in simultaneous operation. The first method

involves implementing the cellular grid, and the second method utilizes digital encoding/decoding [5], [4].

6.1 CELLULAR GRID

Figure 5 illustrates how wireless phones are laid out in hexagonal patterns called cells; this is where the term “cell phone” originated. A large city has hundreds of cellular towers (hexagons). When a cell phone/tower is transmitting/receiving at a particular frequency, to prevent interference from the overlapping of another cell phone, no adjacent hexagonal grid is permitted to use the same frequency. However, because cell phones and towers transmit at lower power levels, any frequencies active in one cell can be simultaneously used in non-adjacent cells. This reuse of frequencies allows more simultaneous cell phone calls and hence a lower individual cost to each user [5], [4].

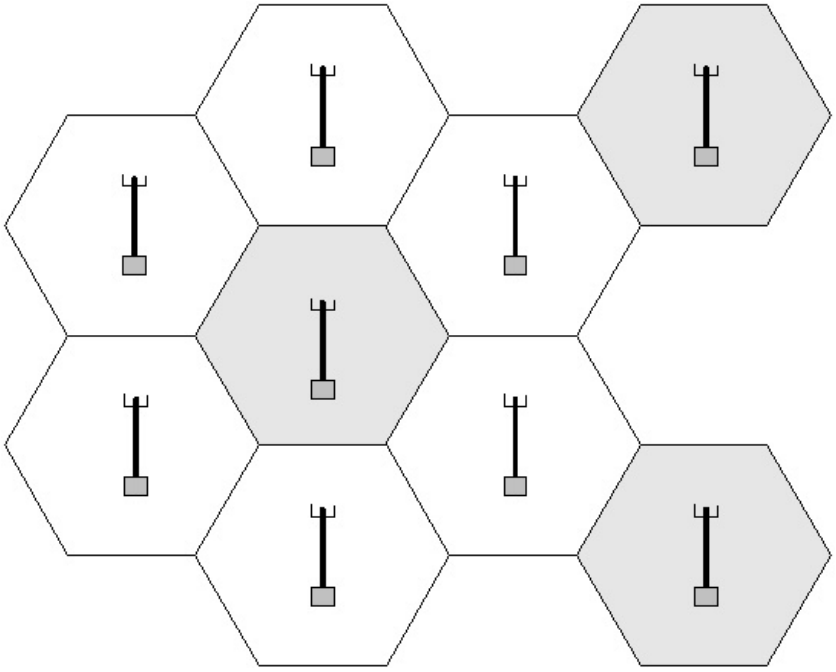


Figure 5: Cellular grid

6.2 MULTIPLE ACCESS

In an analog system, the only method to allow more simultaneous cellular conversations is to increase the density of the towers. However, digital encoding techniques have developed numerous methods for multiplying the number of allowed calls in a cell (i.e. spectral capacity) compared to that of a pure analog system [5], [4]. These techniques are called *multiple access* technologies. The three common technologies are introduced below elaborating only on the details required for the subsequent cell phone interference discussion.

6.2.1 Frequency Division Multiple Access (FDMA)

FDMA divides the FCC-allocated frequency band into narrow-band frequency channels, as shown in figure 6. To achieve duplex (two-way simultaneous) communication the transmitted signal is assigned one channel, while another channel is assigned to the received signal. This simple scheme results in a spectral capacity limited by the allocated frequency band. A pure FDMA implementation is used for analog communication, but FDMA also contributes to the common digital cellular phone standards: IS-54, IS-136, and GSM as described in section 7.1.1.3 [5], [4].

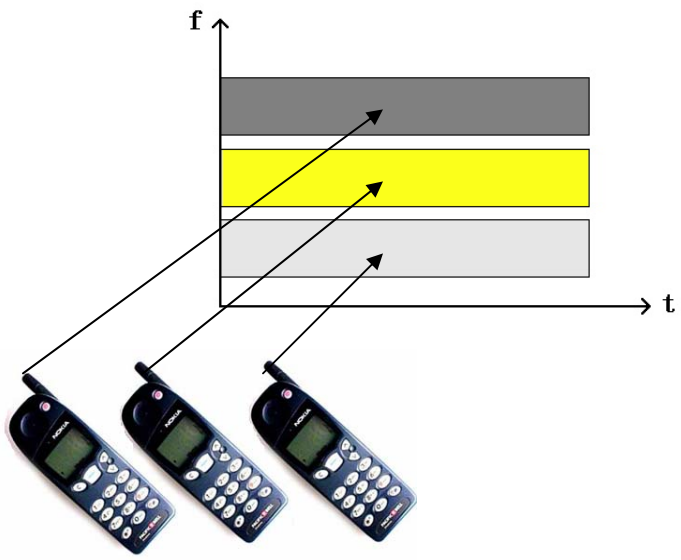


Figure 6: FDMA scheme

6.2.2 Time Division Multiple Access (TDMA)

TDMA multiplies the allowed number of calls per cell by coordinating N cell phones within a cell to share the same frequency. Each cell phone is allotted one time slot transmitting (1/N)th of the time, as seen in figure 7. The real-time outgoing signal is encoded to fit into a time slot of length, slot length (SL), while the incoming signal is decoded resulting in a real-time signal. Each cell phone transmits/receives one time slot per frame [4]. The frame length,

$$FL=N*SL \text{ (sec)}.$$

This scheme results in a large signal (relative to biopotentials) radiating from the cell phone antenna described as bursts of RF information repeating at FL intervals. Each cell phone begins transmitting its corresponding slot at a rate of frame rate,

$$FR=1/FL \text{ (Hz)}.$$

This digital encoding/decoding, although reducing the cost to cell phone users, results in Radio Frequency Interference (RFI) in nearby electronic victims, which effectively adds a low frequency signal of FR Hz (which may lie in the biological frequency band). RFI is one of the major barriers in the adapter design [5], [6], [7], [8].

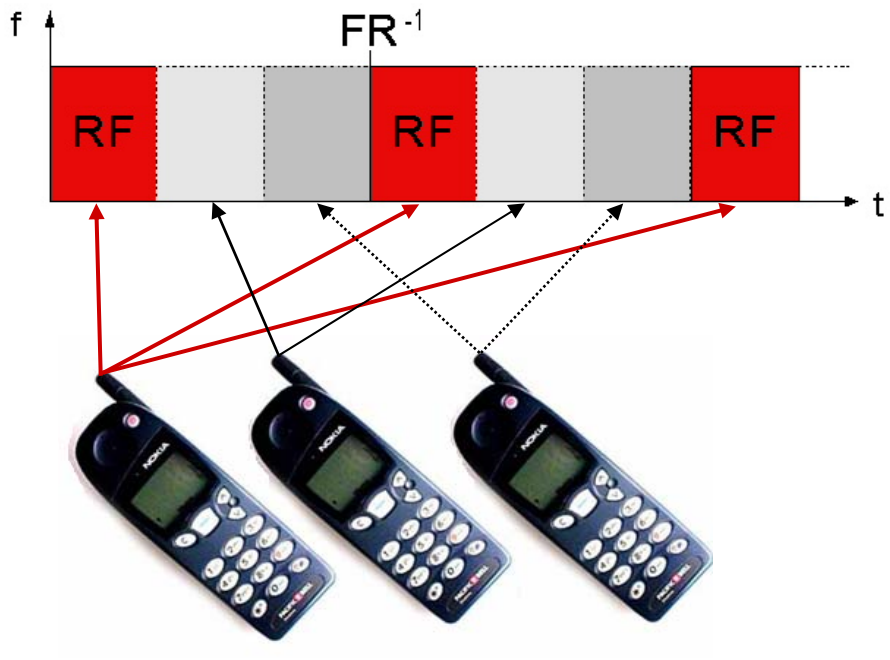


Figure 7: (a)TDMA Transmission, (b) Transmission Signal from each TDMA cell phone

6.2.3 Code Division Multiple Access (CDMA)

To achieve higher interference immunity, CDMA spreads the signal by transmitting small discrete packets across the entire FCC-allotted frequency band. A unique code is assigned to each cell phone, transmitting continuously & simultaneously on the entire frequency band, as shown in figure 8 [6], [4]. Because each cell phone transmits continuously there is no low-frequency behavior, such as the TDMA frame rate, that can be "demodulated" to the biological frequency range. However, as discussed in the Interference section, CDMA can still affect the amplifier's DC output degrading the amplifier's linearity or possibly saturating the signal [4], [5], [6].

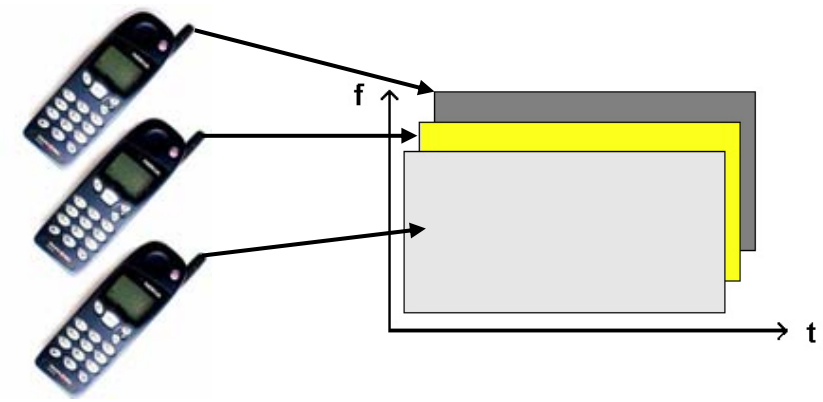


Figure 8: CDMA scheme

6.3 CELLULAR STANDARDS

Table 1 is inserted below to introduce the four most popular standards in the U.S. while highlighting that each standard is a conglomerate sub-standard (each row).

Table 1: Cellular Standards in the U.S. [4], [5],[9]

Standard	AMPS	TDMA (IS-54/IS-136)	GSM	CDMA (IS-95)
Mobile Frequency Range (MHz)	824-894	824-894	1850-1910	824-894
		1850-1990		1850-1990
Multiple Access Technology	FDMA	TDMA/FDMA	TDMA/FDMA	CDMA/FDMA
Peak Transmission Power (W)	3	0.6	1	0.2
TDMA Frame Rate (Hz)	N/A	50	217	N/A
# of Channels	832	832	124	20
Users per Channel	1	3	8	15-50
Channel Spacing (kHz)	30	30	200	1250
Channel Bit Rate	N/A	48.6 kb/s	270.833 kb/s	1.2288 Mb/s

6.3.1 Advanced Mobile Phone System (AMPS)

The AMPS standard employed purely FDMA technology operating in the 800 MHz band. It is now a dead standard but it is include in the table for comparison with the digital technologies. Without any digital multiple access schemes, each cell providing AMPS service is limited to only about 56 voice channels; each conversation uses two of the allotted 832 channels, with 42 being reserved as control channels [4].

6.3.2 Time Division Multiple Access (TDMA)

TDMA is the accepted nomenclature for the Interim Standard 54 (IS-54) and the Interim Standard 136 (IS-136). The transmission frequency for IS-136 (1.9 GHz) is higher than that of IS-54 (800 MHz). As higher frequencies are attenuated less in free space than lower frequencies (for the same distance) [ref: E&M book], IS-136 cell phones can transmit at lower powers than IS-54 phones. Each other sub-standard (row) is common to both IS-54 and IS-136. These standards employ both FDMA and TDMA. The frequency band is first divided into channels (TDMA). Each channel is then divided into time slots (TDMA). Note that TDMA triples the allowed number of simultaneous conversations per tower, using a frame rate of 50 Hz. [9], [4]. Note that this frequency lies within the standard frequency range of most biological waveform types [2].

6.3.3 Global Systems for Mobile Communications (GSM)

GSM also employs a hybrid technology comprised of FDMA and TDMA, first dividing the allotted band into channels (FDMA) then multiplying the allowed number of channels by eight (TDMA). GSM utilizes a wider band per channel to accommodate encryption for increased security and high-bandwidth applications (e.g. internet access). This results in a lower total of allowed calls [5], [4]. Note that the frame rate is 217 Hz, existing above the frequency range for most biological waveform standards [2].

6.3.4 Code Division Multiple Access (CDMA)

As Interim Standard 95 (IS-95), known as the CDMA standard, does not employ TDMA, but rather transmits continuously, there exist no low frequency characteristics to be demodulated to the biological frequency range. The higher frequency IS-95 can transmit at lower power, reducing the incident power for nearby sensitive electronics. CDMA incorporates “background noise” reduction, improving signal quality [4], [9], [5].

7.0 INTERFERENCE

The most common interference in patient-connected devices is caused by the electromagnetic induction between the patient and 60 Hz power lines running through the walls, floor, and ceiling [10]. However, we will forgo its discussion as this type of interference and it is only secondary interference (in this application) dominated by the cell phone interference; its solutions have been well documented in [2], [10], [11], [12], [13] & [14].

7.1 CELLULAR PHONE INTERFERENCE

Designing fro adequate immunity against interference is most difficult when the sensitive equipment is a patient-connected device. It is such a daunting task because the input signal is much weaker than the interference, the structure of the patient-connected device blocks all usual routes of protection, and the patient-connected devices often implement life-critical functions [11], [12].

7.1.1 Threat Parameters

It has been proved in [15] and [5] that the degree of interference is determined by phone's maximum transmission power, the source-victim proximity, the multiple access technology, the

frequency of transmission, and the level of immunity of the sensitive electronic equipment. The first three factors are discussed below. The effect of the frequency of transmission is covered in the demodulation section. Subsequently, methods of increasing the immunity (or compatibility) of patient-connected devices are suggested.

7.1.1.1 Source-Victim Proximity

The incident power, P_i, is inversely proportional to the distance of separation between the cell phone antenna and the vulnerable electronic system;

$$P_i=P_t/(4*P_i*d\text{^}2)\ (W)\ [16].$$

This warrants the distance of separation as one of the dominant threat factors [5]. To minimize risk of critical medical device functionality, hospital policies may enforce either a policy banning the use of cell phones in patient areas, or keeping a minimum distance of 2 m between the cell phone and sensitive equipment. The latter is enforced because findings in [7] show that with a separation of 2 m 0.8% of tested medical devices experienced interference from GSM phones and none of the medical devices tested were interfered with by CDMA; TDMA was not tested but would have been worse.

7.1.1.2 Transmitted Power

The previous equation indicates that the incident power is proportional to the transmitted power from the cell phone. This was reflected in [7], when the lower frequency (higher P_t) GSM phone, caused more interference than the higher frequency (lower P_t) GSM phone. A weakened received signal cues the cell phone to increase transmission power. A stronger received signal cues the cell phone to decrease transmission power. If the patient is moving very fast (e.g. 60 mph in a car), the slow rate of output power fluctuation may have similar effects to TDMA frame rate interference [5], [7].

7.1.1.3 Cellular Phone Standard

The TDMA standard's frame rate of 50 Hz has the medical misfortune of lying within the biological frequency range. RFI not suppressed at the input lines may result in signal distortion at the frame rate and each of its harmonics within the biological frequency range. The GSM standard's demodulated frame rate interference of 217 Hz settles outside of the standard frequency range of most biological waveforms. CDMA phones transmit continuously and do not produce parasitic low frequency harmonics. Even though GSM and CDMA phones do not produce interference within most biological frequency ranges, all three digital standards are capable of shifting the DC operating point of the amplifier's output degrading its linearity or saturating the signal, by the same phenomenon discussed below [17], [5], [15] & [18].

7.1.2 Demodulation

Radio frequency signals existing at the inputs are not present at the output of the bioamplifier, but can generate low frequency interference. The low-frequency spacing of the transmitted RF (i.e. frame rate in TDMA technology) is demodulated lying within the biological frequency range [17].

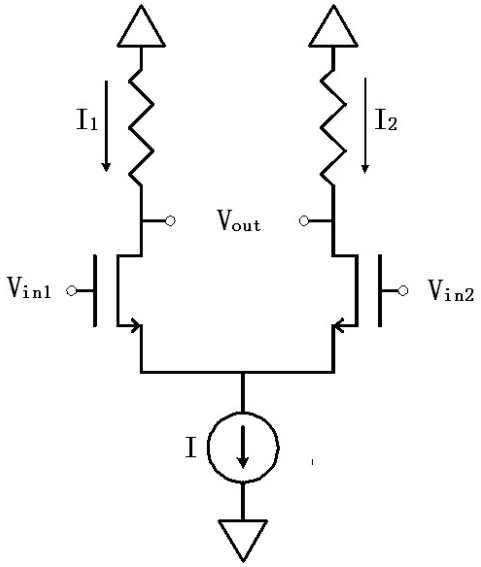


Figure 9: nMOS differential pair

The differential pair is employed to achieve differential amplification while rejecting any influence by the common-mode signal,

$$V_{in,cm} = (V_{in1}+V_{in2})/2.$$

This is accomplished by biasing both input transistors with the same current source, I, to make the output currents I_1 and I_2 independent of $V_{in,cm}$. The current source's output impedance is a

measure of its ability to remain constant. With an ideal current source the output impedance is infinite; therefore connected circuitry can have no effect on I [19]. However at high frequencies the output impedance of the current source greatly decreases, due to its large parasitic capacitance, allowing I to fluctuate when the pair is perturbed by high-frequency $V_{in,cm}$. Assuming I to now be a variable, and expanding the differential pair's voltage-to-current equation as in [18], it can be shown that the resulting differential current output is dependent on both the differential-mode input,

$$V_{in,dm}=V_{in1}-V_{in2},$$

and the product of $V_{in,dm}$ and I (which, at high frequencies, is dependent on Vin,cm). This effect of RF is a resulting DC-shift in the differential output current; this same phenomenon is exploited in the design of RF mixers [20]. The RF is not continuous in the TDMA technology, rather it begins (and ends) transmission at the rate of FR (the frame rate), as seen in figure 10(a). Hence the "DC-shift" fluctuates at FR, resulting in interference at FR Hz [15], [4], [5], as shown in figure 10(b).

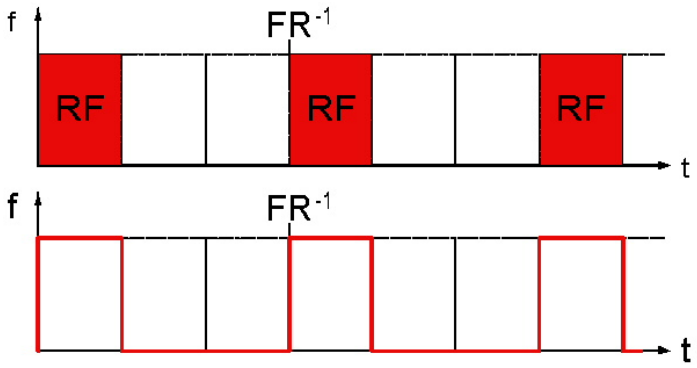


Figure 10: TDMA frame rate demodulation

As the resulting interference is dependent on the product of $V_{in,dm}$ and $V_{in,cm}$ [18], this design approach will aim to reduce both the differential and common-mode RF appearing at the inputs. Because [17] found that BJT op amps are more susceptible to this type of interference than the MOSFET op amps, the design should only incorporate MOSFET amplifiers.

8.0 ELECTROMAGNETIC COMPATIBILITY

Designing for Electromagnetic Compatibility (EMC) is the act of making an electronic device invulnerable to electromagnetic interference, EMI (i.e. compatible with a large EMI environment). EMC requires reducing the interference below a required limit when operating in the harshest electromagnetic environment [13]. EMI can usually be adequately diminished by reducing loop areas, shielding, filtering, proper grounding, or restricting cell phone use. However, the basic rules of EMC are incompatible with patient-connected devices for numerous reasons [14]. As the interfering source is required for communication, it cannot be shielded or appreciably distanced from the sensitive system. The sensitive system includes the patient; the patient cannot be practically shielded. Since biopotential monitoring usually entails differential measurement between two widely spaced electrodes, we are unable to effectively reduce magnetic field interference from twisting the input wires; according to Lenz's Law, decreasing the conductor's loop area will reduce the current magnetically induced [16]. The shielded cables cannot be grounded at the patient end; the electromagnetically-induced current in the shield traversing the resistance of the shields results in a voltage potential between the shield and the input cables at the patient end [14]. This potential difference couples to the signal line via the parasitic capacitance between the shield and signal, resulting in interference pickup. The input cables are often distanced from one another. If the interfering source is nearby, one input is significantly closer to the interfering source than the other; this results in more interference existing on one line than the other, creating a differential RFI. As discussed in the previous

section, the resulting nonlinear effects are dependent on the product of the differential-mode and common-mode RFI [18]. Also, filtering is a tradeoff, in patient-connected devices. Using a capacitor to reduce the shunt impedance at higher frequencies (to reduce RFI) inevitably reduces the input impedance in the desired frequency range, reducing the effective CMRR [14]. Hence patient-connected devices often require a defense from numerous facets of attack at all levels of the design process. This can usually not be achieved with patchwork in the last stages of design. In addition, our design requires the near operation of a cell phone, which has been previously described as a notorious culprit in numerous medical electronic malfunctions.

By definition of an electrical circuit (currents only flow in closed paths), in order for an interfering current to exist, it must originate and return to its source forming a complete loop. The succinct goal of the EMC techniques of shielding, filtering, and grounding is to identify this path and to provide the most direct and low impedance return path for the interference to return to the reference of its interfering source [13].

8.1 CELL PHONE AND ADAPTER PLACEMENT

As mentioned in section 7.1.2, the resulting demodulation is proportional to the product of the common-mode and the differential-mode RFI [18]. The cell phone can be positioned to reduce the differential-mode RFI, RFI_{dm}. By simply clipping the cell phone to the patient's back pocket, the direct path of RFI is shielded by the conductive nature of the patient, producing mostly common-mode RFI, RFI_{cm}, reducing RFI_{dm} at the inputs. Most RFI that reaches susceptible

(unshielded ends of the patient-connected leads) arrived via an indirect, longer path causing the interference on both lines to be mostly common. Hence, positioning the cell phone and the adapter (and cables) reduces the RFI_{dm}, reducing the RFI_{cm} and RFI_{dm} product, thereby reducing the cell phone interference. To minimize the conversion of RFI_{cm} to RFI_{dm}, the input lines should be closely matched [5].

8.2 SHIELDED CABLES

The capacitance between the interfering source and the input leads generates a displacement current. As the input impedances of the adapter are very high, I_d flows from the leads via the skin electrodes, through the patient, to earth-ground. With a mismatch in the skin electrodes, the displacement currents produce different voltage drops across the skin electrodes resulting in a potential difference at the amplifier's inputs. Due to the unavoidable mismatch of arm electrodes, shielding the input leads (removing all mutual capacitance between the interfering source and the leads) is the only way to remove this type of interference coupling [12].

Reducing interference coupling is especially important to maintain a respectable SNR at the input lines containing extremely low amplitude signals. The output lines are less sensitive to interference since the signals have already been amplified and hence have a higher SNR [13]. Also in our application the cell phone and mains interference do not lay within the phones audio band of 400-3400 Hz, and would hence be filtered out by the phone's band-pass filter prior to the phone's analog-to-digital conversion.

The dominating interference is generated by the cell phone transmission. The reference of the source of this interference is the negative terminal of the cell phone's battery which will be defined as "ground." To minimize the effects of this interference, the designer should provide the most direct, lowest impedance path to the interfering source's reference as possible. This can be achieved with shielding by connecting the shields to the ground plane [13], [14].

8.3 CHASSIS

Enclosing the circuit within a conductive chassis electrically isolates the internal circuit from external EMI (except for low frequency, predominantly magnetic interference) [13], [14]. When confronted with the increased cost, size, and weight of the adapter, and considering the fact that we have other EMC options, and this is not the dominant interference, the chassis becomes an expensive and unnecessary means of EMC.

8.4 RFI FILTERING

Unless proven necessary, filtering at the input lines is often undesirable and avoided as it inherently decreases the input impedance while creating more impedance mismatching, thereby decreasing the effective CMRR [17], [10]. However, as previously discussed, nonlinear voltage-to-current conversion characteristics of the input operational amplifiers demodulate the RFI down to its burst frequency (frame rate, FR), which lies within our biological waveform

frequency range at larger amplitudes. Since, we can not shield the cell phone transmission signal to reduce the source of interference, and we cannot shield the patient along with the cables and adaptor to isolate the entire circuitry, we found it necessary to reduce the interference via low pass filtering at the inputs.

By designing the filter to have a cutoff frequency as high as possible (i.e. just below the fundamental frequency of interference) we can best preserve the CMRR for the lower frequencies of interest, keeping the input impedance for lower frequencies as high as possible. . By using a higher order filter we can achieve a steeper roll-off, allowing us to achieve more filtering with a higher cutoff frequency. To avoid drastically increasing the series resistance (when cascading multiple filters), while increasing the order, LC filters should be employed [21]. The filter should achieve sufficient attenuation at the interfering frequencies. However the parasitics of a capacitor generate a series resonant frequency, above which the capacitor appears inductive. The parasitics of an inductor generate a parallel resonant frequency, above which the inductor appears capacitive [22]. To simplify the design, a prefabricated filter should be chosen with its maximum insertion loss (an indication of the degree of attenuation provided by the filter) at the cell phone technology's transmission frequency. The capacitor-to-ground connections of the filter should be as close to the ground plane as possible to reduce the "to-ground" line impedance (due to L at HF [23], [16]), increasing the impedance path to ground, effectively amplifying the influence of the EMI [13], [23].

8.5 GROUNDING

A ground loop is a mechanism of conducting unwanted feedback illustrated in figure 11. Component A may be connected to ground via a wire of nonzero impedance at node X. Hence, according to Ohm's law (v=i*z), the actual voltage at node X, V_X, is equal to the current drawn from component A multiplied by the impedance of the wire to ground. If component B (drawing negligible current) is also connected to node X, sharing the same wire to ground as component A (to save this length of wire), component B will be referenced to V_A, rather than the intended ground. This effectively couples component B to component A.

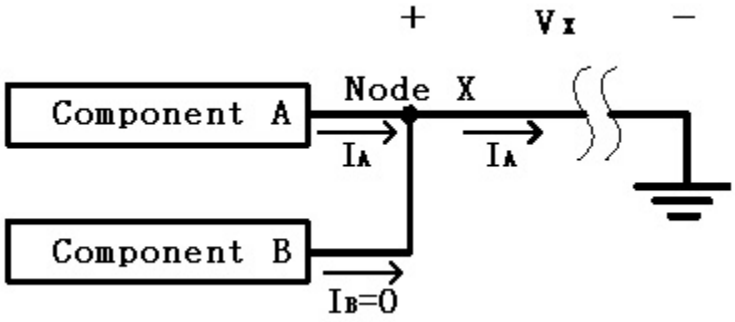


Figure 11: Ground loop between couples B with A via V_X

To eliminate ground loops, the designer must theoretically use separate wires to reference a single point ground [23]. Practically, this could result in large loop areas making the wires prone to magnetic pickup or emission. The most feasible solution is to connect all nodes intended to reference ground to a ground plane with as short and fat of a wire as possible. This is necessary when high frequencies are present because at high frequencies we must model the effective inductance of all wires. This inductance is related to the wire's length-to-width ratio [24], [16].

Hence, short fat wires and ground planes provide a low inductance and hence a low impedance even at high frequencies. The lower the impedance of the ground connections, the less effective the ground loops are at causing interference [23].

8.6 DSP

Experimentally forming numerous combinations of the EMC techniques mentioned above (and many not worth mentioning), I formed the conclusion that the interference from TDMA standard cell phones could not be sufficiently prevented. The frame rate of the TDMA standards (50 Hz) exists within most biological frequency ranges. The FR interference could not be reduced to less than 1% of the biological waveform peak-to-peak amplitude, the SNR standard [12]. Even if my efforts were successful, the EMC solution would not have been cost, space, or power efficient. Instead, I postulated that complete EMC would have been in vain. To keep this adapter design efficient, it was only necessary to reduce the resulting interference low enough to allow me to amplify the signal high enough to achieve the desired SNR (40dB) without saturating the microphone's input (limited by 45mVp-p on our model).

As the interference occurs at well-defined pin-point frequencies (FR and its harmonics), there exist many digital signal processing (DSP techniques to remove the interference resulting in minimal signal distortion. I chose to compute the Fast Fourier Transform (FFT) in MATLAB, subsequently removing the Fourier series coefficients at the contaminated frequencies. These

frequencies can be removed with stop-bands of +/- 0.1 Hz. These coefficients can be replaced with the average of the adjacent remaining coefficients, resulting in minimal distortion.

9.0 LOW NOISE DESIGN

As we will be interfacing the cell-phone via the microphone input which expects signals of very low amplitude (for our model ~45mV_{p-p}), we must design this adaptor as a low noise device to result in a respectable SNR. There are three sources of noise in electronics: resistor, direct current and active devices [23].

The noise level in bioelectric measurements should be kept below 1% of the peak-to-peak biopotential signal (i.e. 40dB) [12]. Properly designing a circuit, for low noise constraints, often requires making the active devices the dominant noise source (not the resistors) [23]. So, the design should begin by choosing all ICs to be rated for noise voltages (around 10Hz) to be less than 1% the maximum allowed biological waveform of 45 mV_{p-p}, as constrained by the cell phone's microphone input. All external resistors were then chosen to maintain the active noise voltages as the dominant noise source by keeping the voltage seen at the inputs of an active device due to the noise current flowing across the resistors less then or equal to 10% that of the active noise voltage. To minimize power dissipation this design utilizes the maximum resistances which obey the active noise constraints. It was then verified that the resulting resistors did not dominate the noise with regards to passive thermal or shot noise.

10.0 SHARING THE CELL PHONE BATTERY

In the interest of minimizing the size and weight of the adapter, the cell phone's battery can be used to power the adapter, eliminating the weight, required-space, and replacement of additional batteries. We can access the battery terminals of all cell phones, as such ports are required to recharge the cell phone battery. However, during operation (especially transmission) the cell phone's power supply voltage varies greatly making its direct connection, not a desirable power supply. To stabilize the supply voltage, a voltage regulator should be employed [23].

11.0 MODULATION

As the frequencies comprising biological waveforms can not pass through the communication channel, we must modulate our signals frequency content [24]. It is known that multiplication in the time domain corresponds to convolution in the frequency domain. This is why multiplying an input signal with a sinusoid effectively shifts the input signal's frequency content to the frequency of the sinusoid. However, modulating the input signal's frequency content using a square wave can be shown to have a more power and space efficient implementation. Figure 12 illustrates that multiplication of the input signal with a square wave results in replicas of the input signal's frequency content being shifted and scaled according to the Fourier coefficients of the square (carrier) wave. This corresponds theoretically to a classical modulation technique known as *Amplitude Modulation with a Pulse Carrier* with a pulse duty cycle of 50% [24].

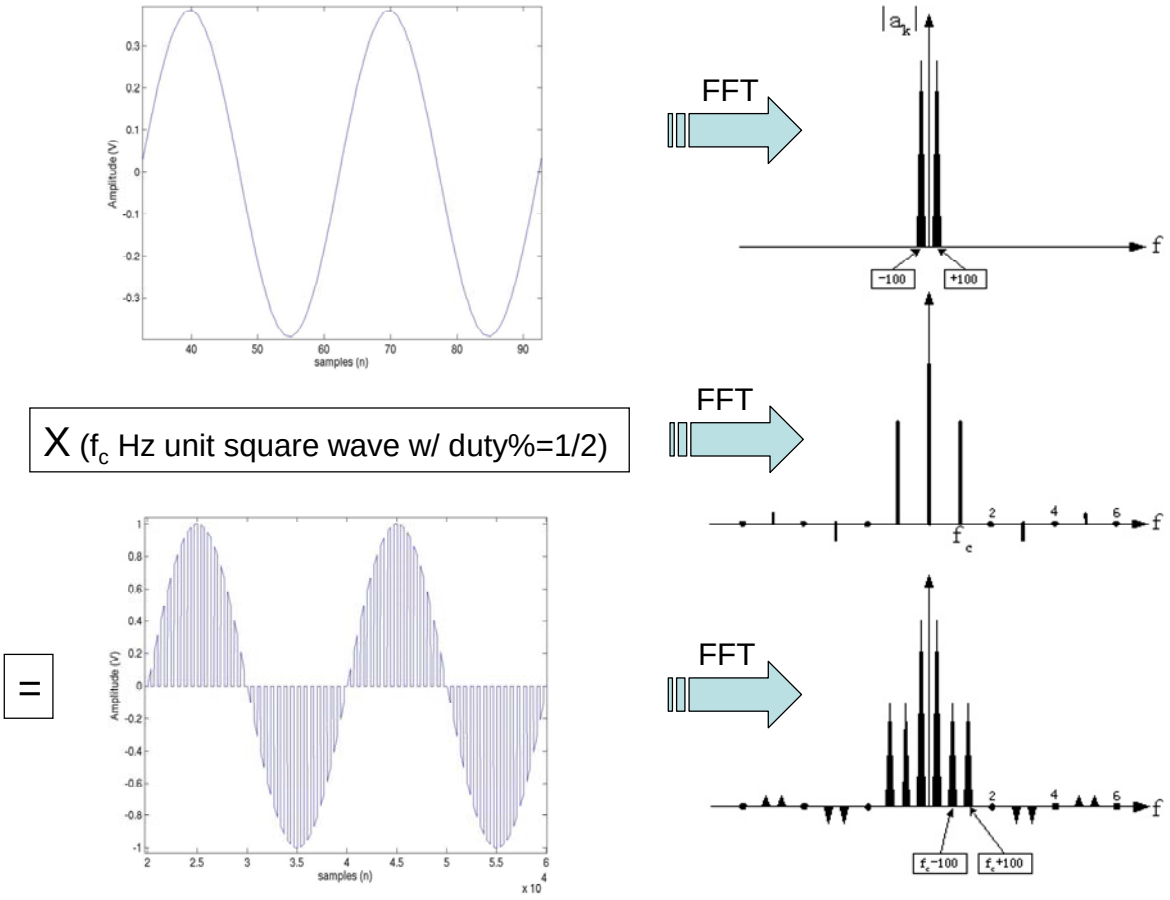


Figure 12: Amplitude modulation with a pulse carrier with duty%=50

Multiplication with a unit square wave can be implemented with a solid state single-pole double-throw (SPDT) switch, as shown in figure 13. By periodically switching the output from the input signal to its reference, we generate time-slices of the input signal at the output at a rate and duty cycle dictated by the clock. This is mathematically equivalent to multiplying the input signal by a periodic sampling pulse and can therefore be used to implement *Amplitude Modulation with a Pulse Carrier* [25].

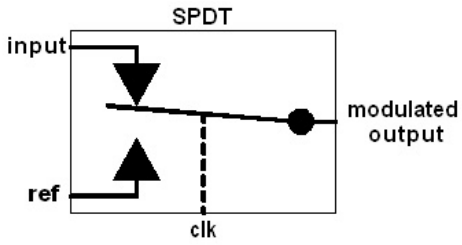


Figure 13: Modulation implementation

An example using a 100 Hz sinusoid as the input signal is illustrated in figure 14. In the Fourier domain, shown in figure 15, the resulting output contains replicas of the input signal at its original frequency as well as harmonics of the switching frequency (3 kHz) which are scaled according to the carrier's duty cycle (50% for a single input signal).

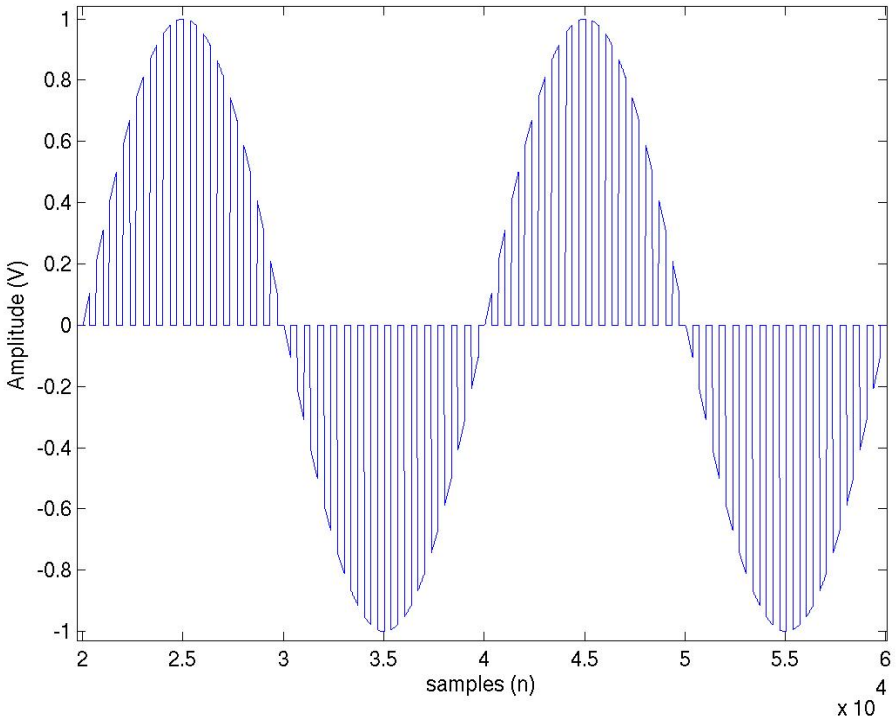


Figure 14: Product 100 Hz sinusoid and 3 kHz pulse, simulating the figure 13 implementation & resulting in modulation

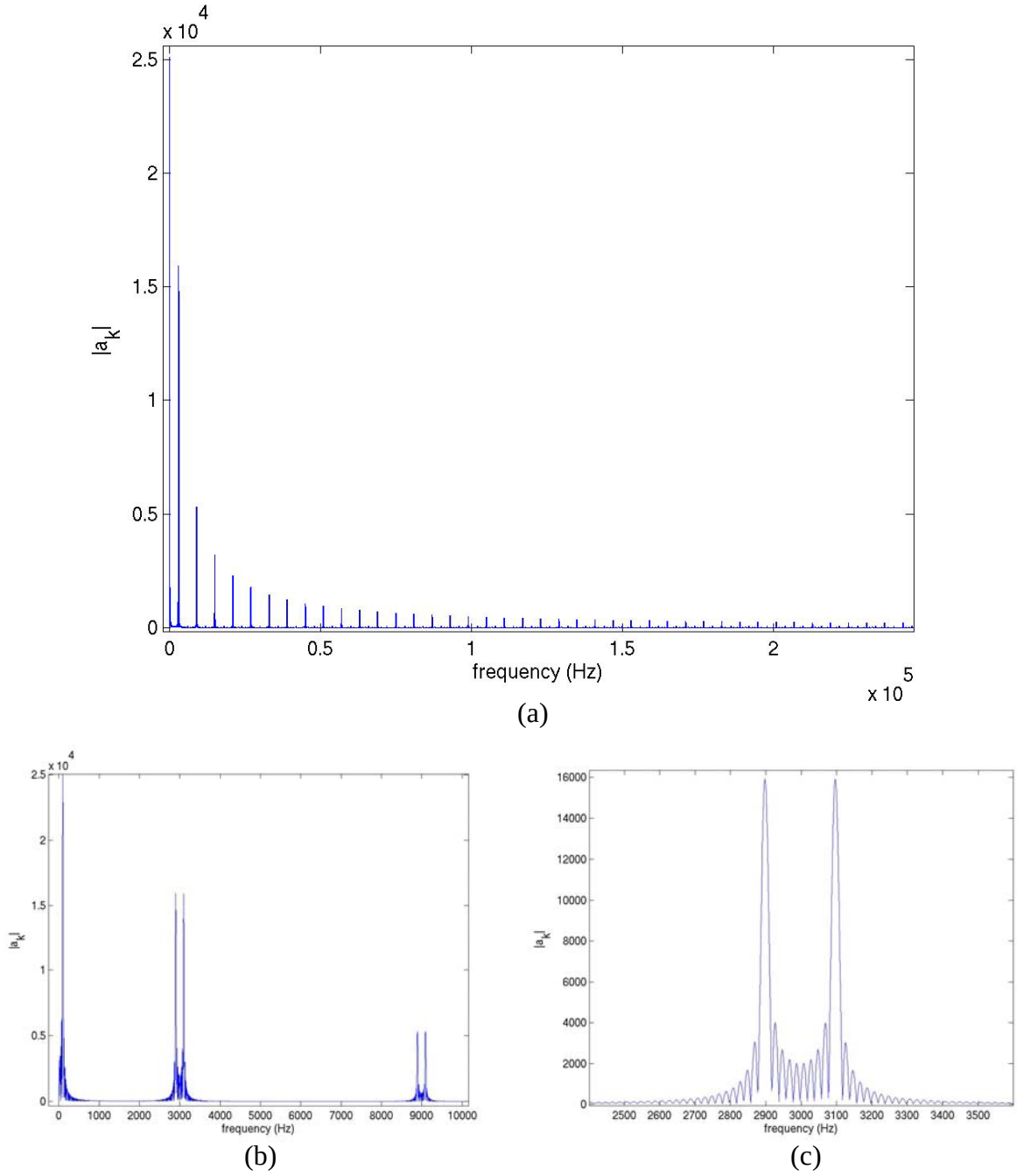


Figure 15: (a) FFT of modulated sinusoid in figure 14 zoomed in (b) 0-10 kHz (c) 2-4 kHz

Upon analysis of the Fourier coefficients in figure 15, it is shown that the original signal frequency still exists in the modulated output and that it is larger than all the modulated

components. To allow us to increase the modulated signal gain (improving SNR) without saturating the microphone input, we can filter the larger original low-frequency components and all harmonics of the carrier frequency, as shown in figure 14 with the corresponding FFT in figure 15. By decreasing the range of the input signal, we can further increase the gain of the amplifier increasing the amplitude of the signal to be passed through the phone system (400-3400 Hz) to improve the SNR.

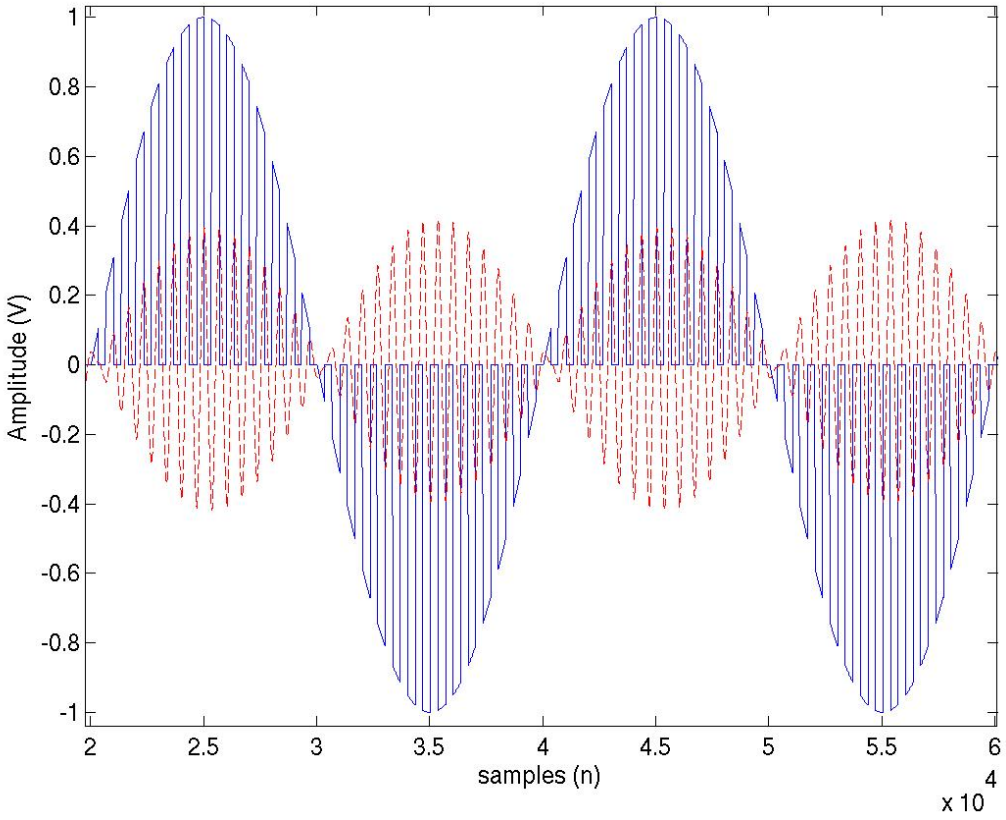


Figure 16: Modulated sinusoid Band-pass filtered between 400 and 3400 Hz

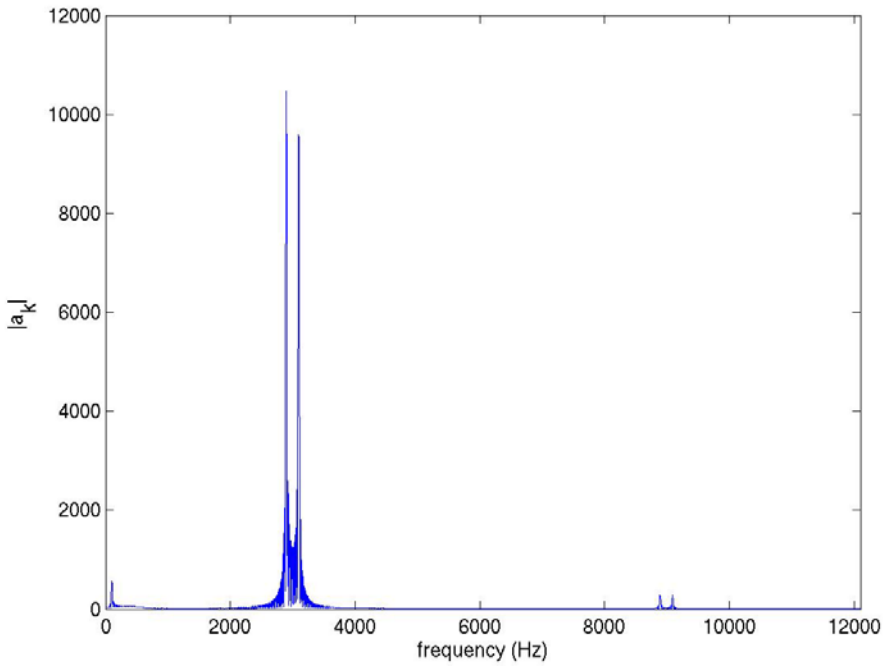


Figure 17: FFT of band-pass filtered sinusoid

12.0 DEMODULATION

We can detect the carrier frequency of the received modulated signal by computing the Fast Fourier Transform in MATLAB. If there is a large DC offset at the input then the carrier frequency will be the point of the maximum peak in the range of the user-entered approximate carrier frequency. If the offset is low (as in figure 14), then we can calculate the carrier frequency by band-pass filtering around the approximate carrier frequency then detecting the largest peak on the left-hand side (LHS) and the right-hand side (RHS) of the approximate carrier frequency (see figure 17). Due to the symmetry of the input signal about the carrier frequency with the LHS corresponding to the negative frequency component of the input signal, and the RHS corresponding to the positive frequency component [24], we can acquire a precise measurement of the carrier frequency by averaging the frequencies at which the RHS and the LHS peaks occur [25].

Figure 18 demonstrates our demodulation technique. After acquiring the first peak in the received signal (i.e. the modulated output, post high-pass filtering), we can demodulate the received signal by plotting each multiple of the time period of the carrier signal (i.e. the inverse of the carrier frequency) at times referenced to the first peak occurrence. The demodulated signal would be a subsequent plot of the sampling points represented by the circles in figure 18, forming an accurate representation of the original input signal, seen in figure 19.

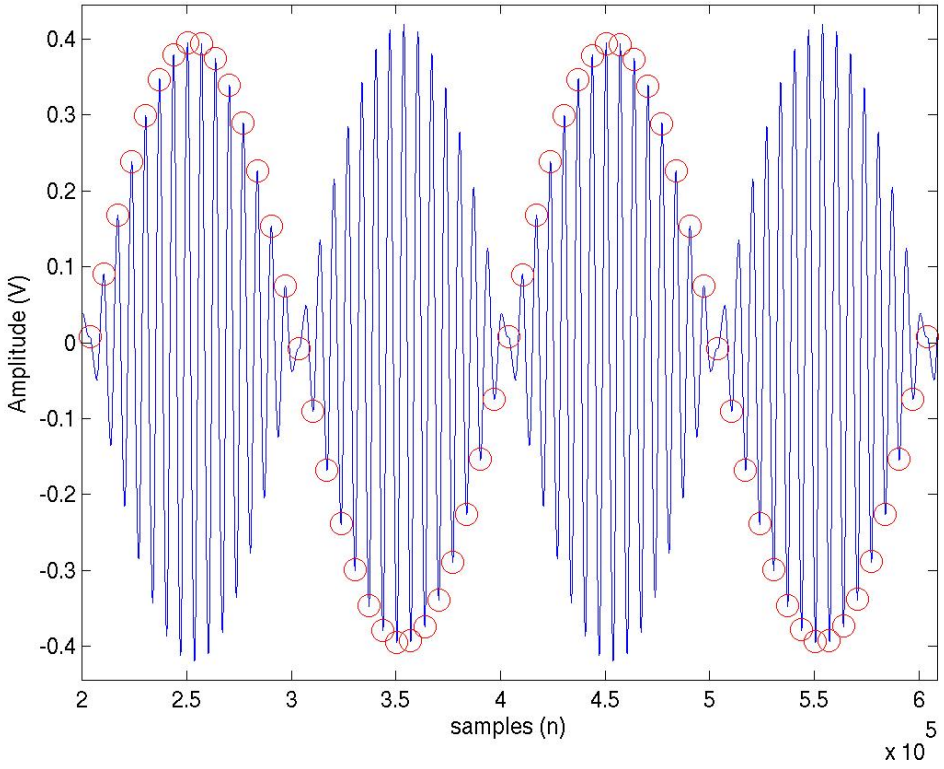
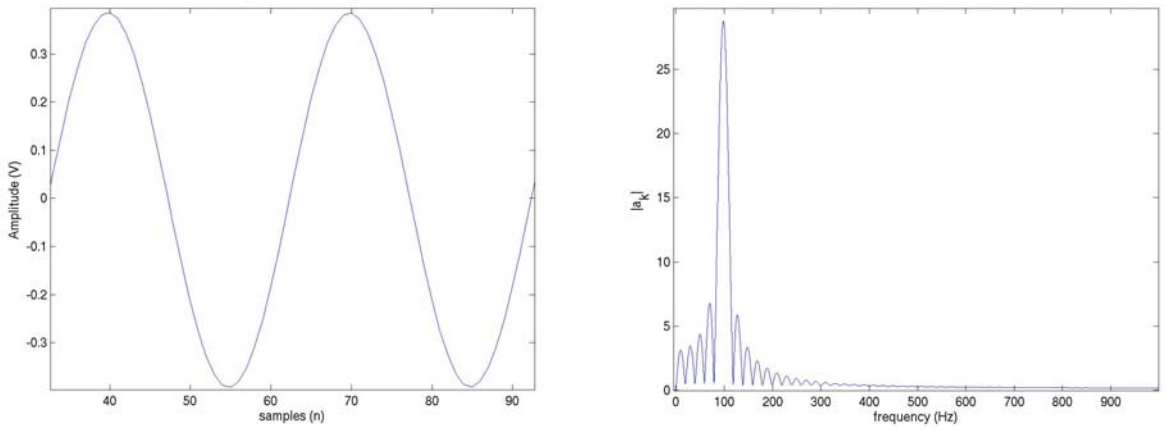


Figure 18: Input signal data points (circles) recovered from band-passed modulated signal (solid line)



(a) (b)

Figure 19: Representation of original 100 Hz sinusoid in the time domain (a), and the frequency domain (b)

13.0 CELL PHONE INTERFACE

In order to keep our application practical, we can not modify the cell phone. Rather we interface the biological inputs with the cell phone using an adapter. For this adapter design to be attractive, it must be general compatible with the majority of cell phones in use today. The most common data port for cell phones is the *earbud*, which makes cell phone communication, while driving, safer (and sometimes legal), by combing a speaker which sits in the user's ear, allowing an omni-directional microphone to dangle by the user's mouth. To reduce size, weight, and battery replacement, 90% of these microphones are two-wire electret microphones which require *plug-in* power for a very small preamplifier [26]. In this manner, the same wire used for data communication is also employed to provide power to the onboard preamplifier. The most common interface for these microphones is shown below in figure 20 [26].

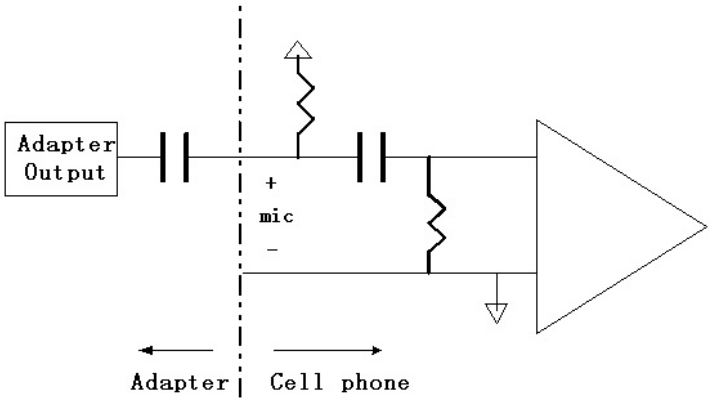


Figure 20: Interfacing with a port providing "plug-in" power for a microphone

Directly connecting the modulated output to the microphone port on the cell phone would try to bias the output close to the battery's voltage (4.8V for our cell phones). However, the max voltage range of the adapter's circuitry is between the cell phone's ground and the regulator's output voltage (~3.3V). Therefore a direct connection to the microphone input would saturate circuit's output signal. The simplest solution is to AC couple the adapter's output to the microphone input by connecting the two with a series (DC blocking) capacitor. The capacitance value can be chosen to combine with the cell phone's load impedance to form a high-pass filter, removing the dominant low-frequency signal content; these would be subsequently filtered by the phone prior to transmission but may saturate the microphone's input stage (with a small dynamic range ~45mVp-p). If the low-frequency content is filtered within the adapter, then the designer can increase the bioamplifier gain increasing the signal-to-noise ratio.

14.0 LANDLINE PHONE-TO-COMPUTER INTERFACE

To keep the design as simple and safe as possible, one should leave the phone box connected to the wall. The phone box provides surge protection, and it decouples the outgoing signal from the incoming signal. The phone box sends two wires to the microphone and two wires to the speaker of the handset [27]. As this design will only receive signals from the landline phone, one can take the two wires intended for the speaker and couple the signal to an input channel of the computer, such as the audio input of the sound card, sampling at >8kHz as a .wav file. Since the phone system biases the signal around -28Vdc (to prevent copper wire oxidation), one should couple the phone box to the computer using a telephone transformer (specifically designed for the frequency range of interest) to provide adequate isolation, as illustrated in figure 21.

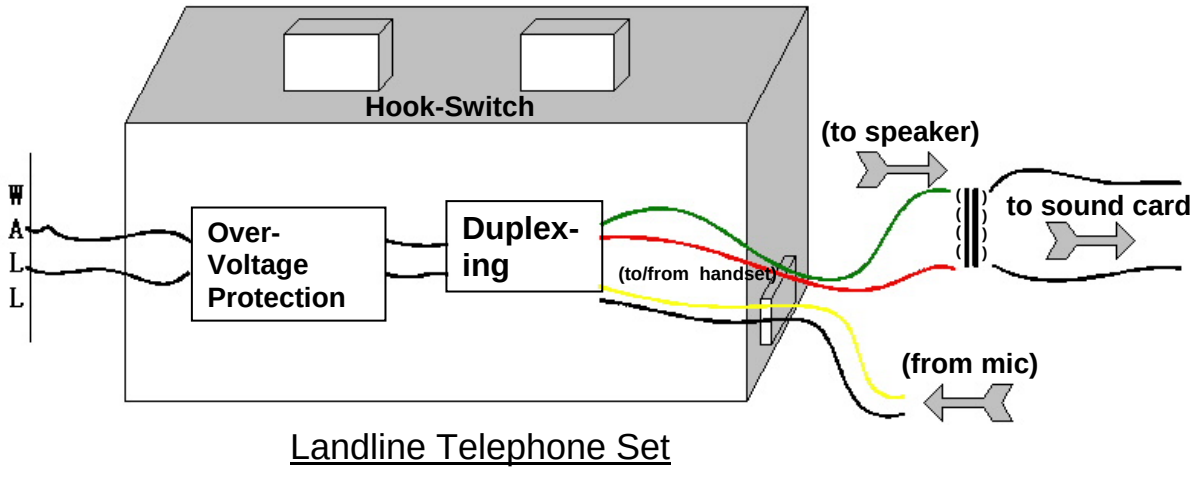


Figure 21: Acquiring data from a landline phone

15.0 SINGLE-CHANNEL ECG TRANSMISSION EXAMPLE

In order to substantiate the ability of cell phones to transmit biological data, this section steps through an experiment where I transmitted one channel of ECG using a IS-54 (FR=50 Hz) cell phone, a sample of which is shown in figure 22. The transmitted and acquired signal is shown in figure 23. The carrier frequency was found by averaging the peaks to the LHS and RHS of the user-entered approximate carrier frequency (as described in section 12).

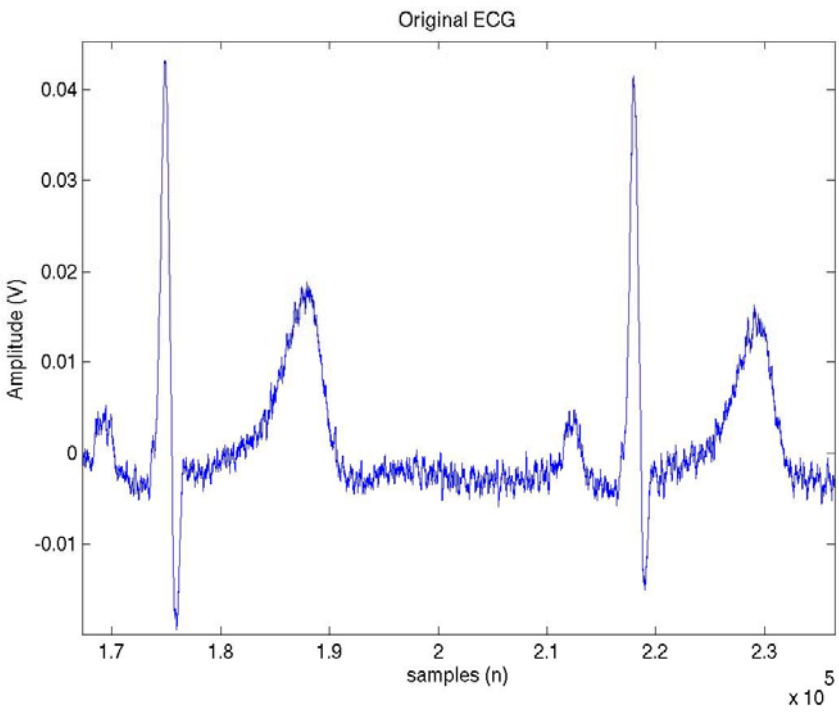


Figure 22: Sample of the original ECG

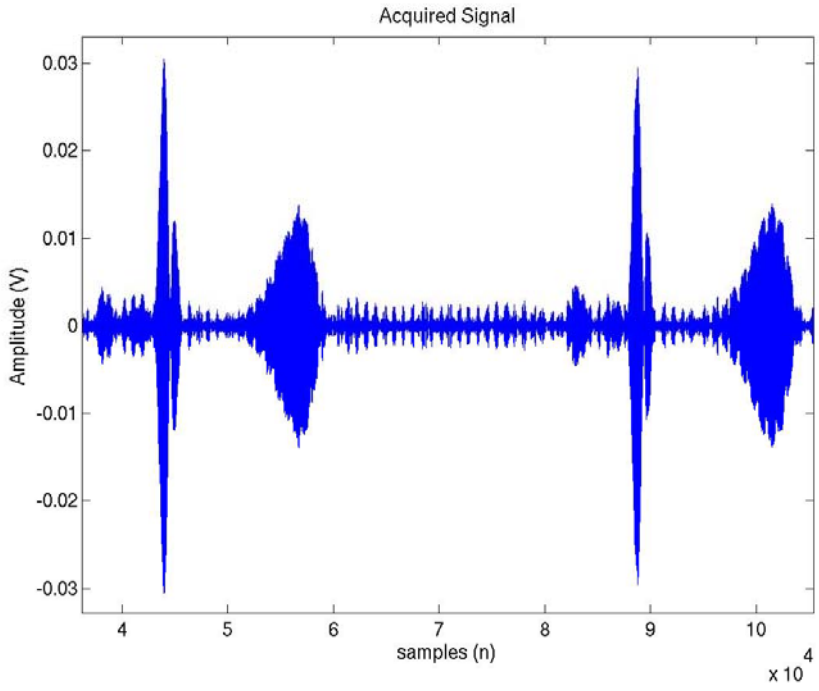


Figure 23: Transmitted/acquired data

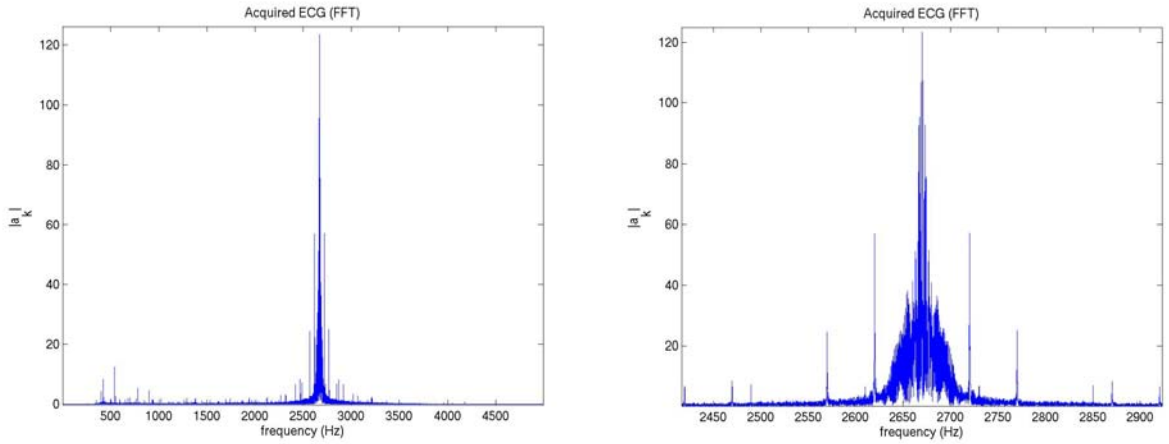


Figure 24: (a) FFT of transmitted/acquired data. (b) Zoomed in.

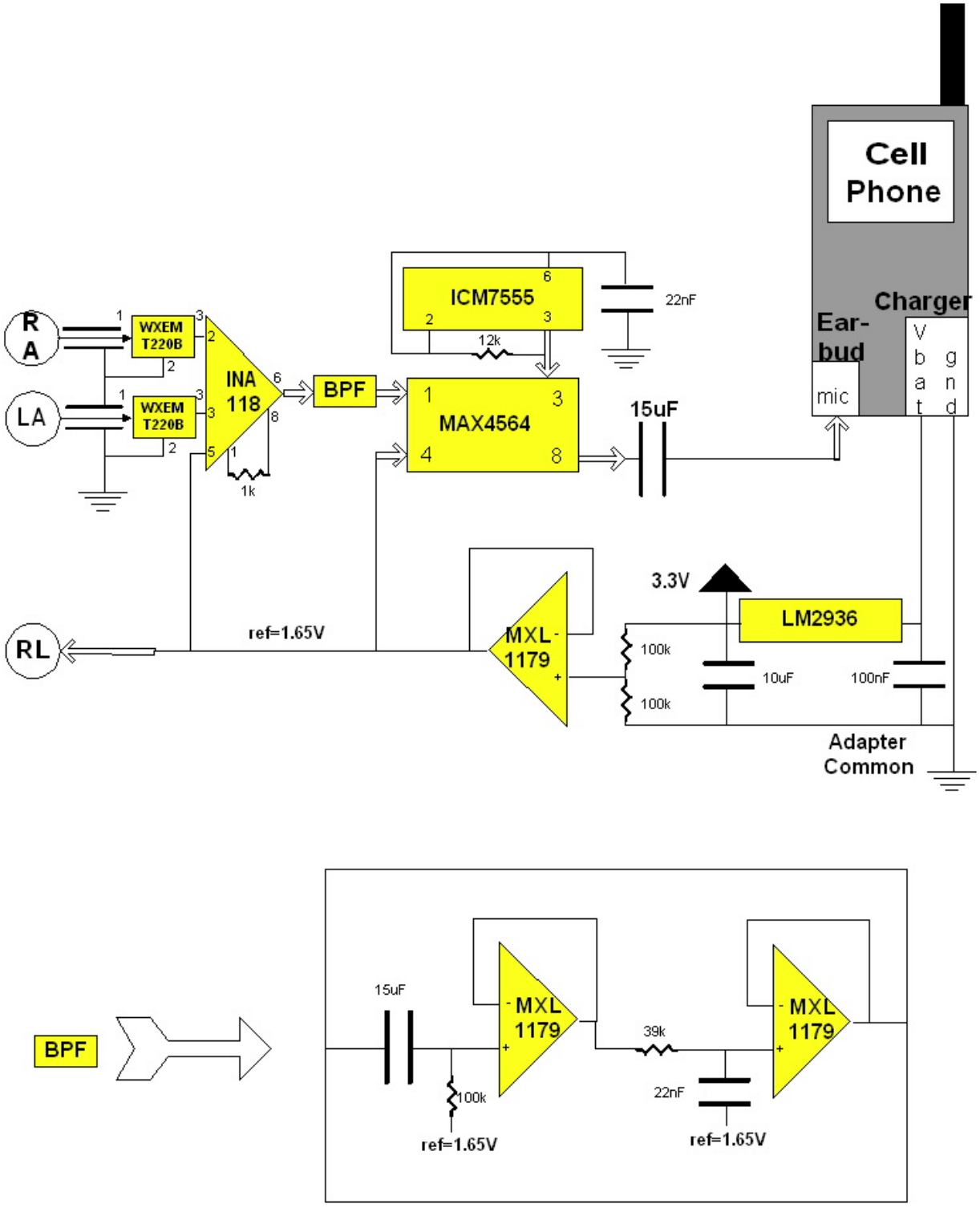


Figure 25: Schematic of adapter used in the example

The first peak in the acquired signal was found and taken as the first point in the demodulated signal represented as the circle in figure 26 (b). The signal was demodulated by plotting multiples of the reciprocal of the calculated carrier frequency referenced to the first peak, as shown in figure 26. Figure 27 plots each of these samples.

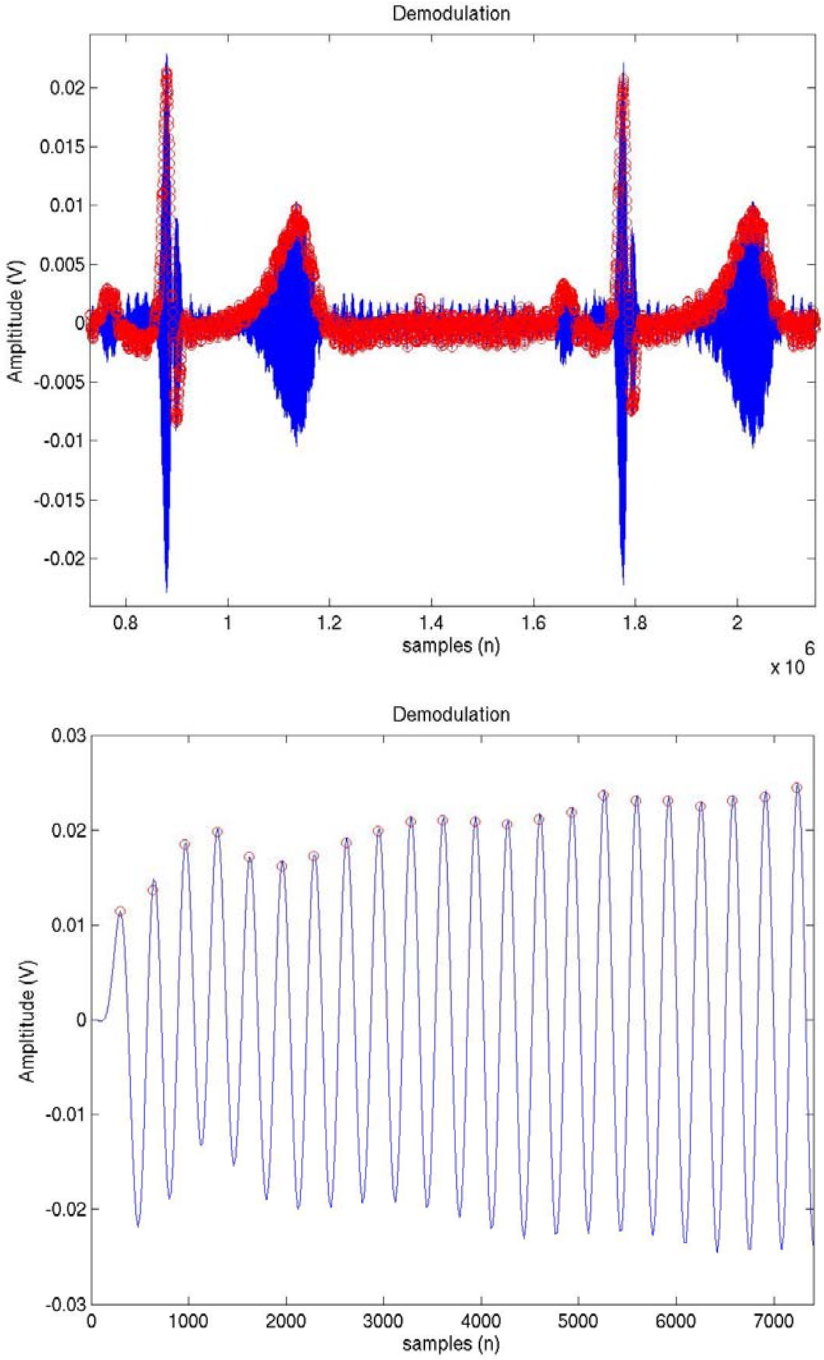


Figure 26: (a) Demodulation of acquired data. (b) Zoomed in.

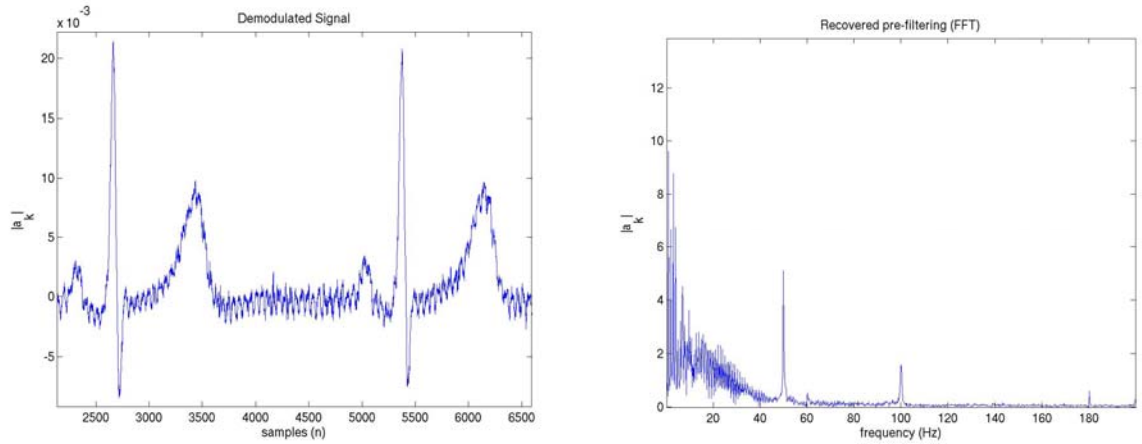


Figure 27: Demodulated data (a) in the time domain (b) in the frequency domain

As this example utilized the IS-54 standard cell phone employing a TDMA frame rate of 50 Hz, the FFT in figure 27 (b) clearly shows the dominant interference is at FR=50 Hz with all of its harmonics. By removing the Fourier series coefficients corresponding to this interference, we can recover a de-noised representation of the original ECG, as seen in figure 28. The m-file used to recover this signal is included in the appendix.

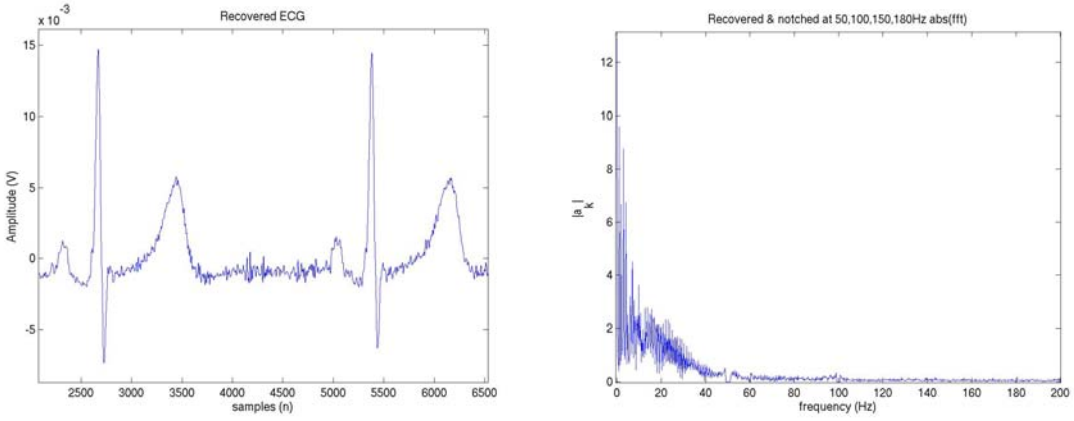


Figure 28: Recovered ECG (a) in the time domain (b) in the frequency domain

16.0 TRANSMITTING MULTI-CHANNEL BIOLOGICAL WAVEFORMS

Often multiple channels of biological waveforms are required for diagnosing patients. Two channels of a biological waveform are required to uniquely describe the bioelectric activity in a single plane [2]. In clinical practice, several channels are recorded. Portable ECG monitoring devices currently on the market provide either one or three channels. When transmitting multiple signals through one communication channel simultaneously, multiplexing is mandatory [2].

The multiple channels would first be multiplexed with a minimum sampling frequency,

$$f_{s,mux} >= 2*BW,$$

obeying the Nyquist rate, thereby permitting signal recovery. Recalling that it is necessary to transmit both the sampling frequency and the carrier frequency ($f_c=f_s*$inputs) to enable demultiplexing, the modulation block must modulate a multiplexed signal with a bandwidth of

$$BW_{mux}=f_{s,mux}*(\text{inputs}).$$

As the phone channel must pass a carrier frequency with the multiplexed signal (with BW_{mux} on each side), it becomes clear that the number of biological waveform inputs is still severely

limited. The phone system can transmit up to six channels of waveforms if limited to BW=85Hz, yet it is remains limited to only two channels with BW=200Hz.

17.0 DISCUSSION

This thesis introduced the necessary background leading to the suggested cell phone technology and the techniques for designing an adapter to facilitate biotelemetry using a cellular phone. By understanding the nature of the cell phone interference an electromagnetic compatible design had been established. A power and space efficient implementation of *Amplitude Modulation with a Pulse Carrier* has been instantiated. The tools were developed to efficiently interface with the cell phone and acquire data from the landline phone. An algorithm to recover a high-fidelity representation of the original waveform was designed and implemented in MATLAB. Transmitting biological waveforms through a cellular phone (following the suggested techniques) has been shown useful and effective.

APPENDIX

BIOLOGICAL WAVEFORM RECOVERY PROGRAM

```
close all;
clear all;

y1=load('nov1_sine100_lidoff_ontable.txt');
y=y1(1:round(length(y1)/10+1));
channels=1; %number of signal channels
refs=1; %number of reference channels
BW=200;
fs=44100; %Hz

samples=length(y); %1 second's worth

duration=samples/fs; %duration of EEG time series in seconds
fc_approx=2670; %user entered approximate carrier clk freq
large_input_offset=0; %1 if true

fc_chosen=2670;


%plot acquired signal
figure(1);
plot(y);
title('Acquired Signal 1s 1ch muxed');
hold on;

%calculate and plot fft of acquired signal
figure(61);
f61=fft(y,1048576);
 fr61=fs*(0:524287)/1048576;
plot(fr61,abs(f61(1:524288)));
xlabel('frequency (Hz)');
title('Acquired Signal abs(fft) ');

%BPF acuired signal around fc_approx to find fc
```

```
[h,g]=cheby1(2,0.5,[(fc_approx-0.25*fc_approx)/(fs/2) (fc_approx+0.25*fc_approx)/(fs/2)]);
figure(117);
subplot(2,1,1);
plot(abs(h));
subplot(2,1,2);
freqz(h,g,512,fs);
grid on;
title('BPF1 freq response to find fc');

%plot BPF1 to acquired signal
BPF1=filter(h,g,y);
figure(115);
plot(y);
hold on;
plot(BPF1,'--r');
title('Acquired Signal (B) Vs. post-BPF1 (R)');

%calculate and plot fft of BPF1 to find fc
figure(116);
f116=fft(BPF1,2097152);
 fr116=fs*(0:1048575)/2097152;
plot(fr116,abs(f116(1:1048576)));
xlabel('frequency (Hz)');
title('FFT to find fc ');

%calculate fc
if large_input_offset == 1
[fft_mag,fft_pt]=max(abs(f116(1:1048576))); %finds my switching freq
fc=(fft_pt-1)*fs/2097152;

else
   fc_approx_index=round(fc_approx/fs*2097152+1);
   [max_LHS_mag, max_LHS_index]=max(abs(f116(1:fc_approx_index)));
   [max_RHS_mag, max_RHS_index]=max(abs(f116(fc_approx_index:1048576)));
   fc_index=round(1000*(max_LHS_index+max_RHS_index+fc_approx_index-1)/2)/1000;
%rounds to the 3rd dec
   fc=(fc_index-1)*fs/2097152;
end

%trying to free memory
clear BPF1 f116 fr116 fr61 f61;


%BPF acquired signal around fc
[r,q]=cheby1(2,0.5,[(200)/(fs/2) (fc*2.5)/(fs/2)]);
figure(14);
```

```
subplot(2,1,1);
plot(abs(r));
subplot(2,1,2);
freqz(r,q,512,fs);
grid on;
title('BPF2 freq response');

%plot BPF2 versus acquired signal
BPF2=filter(r,q,y);
figure(15);
plot(y);
hold on;
plot(BPF2,'--r');
title('Acquired Signal (B) Vs. post-BPF2 (R)');

figure(91);
f91=fft(BPF2,1048576);
 fr91=fs*(0:524287)/1048576;
plot(fr91,abs(f91(1:524288)));
xlabel('frequency (Hz)');
title('post-BPF fft');


%trying to free memory...
clear fr91 f91;


%interpolate BPF2 and find the first sampling point for demodualtion
x_BPF2=1:1:length(BPF2);
x_interp=0.01:0.01:length(BPF2);
y_interp=interp1(x_BPF2,BPF2,x_interp);
fs_interp2=100*fs; %effective sampling freq of interpolated signal
[first_max, start]=max(abs(y_interp(1:round(2*100*fs/fc))));


% this method uses the starting point as a pt ref to eliminate the accumalation
% of estimation error from rounding inc and start and limiting x_interp to 0.1
x1=start;
X1=[];
Ch1ref=[];
X1ref=[];
sample_num=1;
ch1=[];

if channels==1
while x1 <= length(y_interp)
```

```
ch1=[ch1, y_interp(x1)];
X1=[X1, x1];
x1=start + round(sample_num*fs_interp2/fc);
sample_num=sample_num+1;
end

elseif channels==2
   X2=[];
   ch2=[];
   Ch2ref=[];
   X2ref=[];
   x2=start+round(fs_interp2/fc/2);
   x1ref=x1+round(fs_interp2/fc/(channels+refs));
   x2ref=x2+round(fs_interp2/fc/(channels+refs));

   while x2ref <= length(y_interp)
   ch1=[ch1, y_interp(x1)-y_interp(x1ref)];
   Ch1ref=[Ch1ref, y_interp(x1ref)];
   ch2=[ch2, y_interp(x2)-y_interp(x2ref)];
   Ch2ref=[Ch2ref, y_interp(x2ref)];
   X1=[X1,x1];
   X1ref=[X1ref, x1ref];
   X2=[X2,x2];
   X2ref=[X2ref, x2ref];
   x1=start + round(sample_num*fs_interp2/fc);
   x2=x1+round(fs_interp2/fc/2);
   x1ref=x1+round(fs_interp2/fc/(channels+refs));
   x2ref=x2+round(fs_interp2/fc/(channels+refs));

   sample_num=sample_num+1;
end
end

recovered_samples=sample_num-1;
recovered_duration=duration-(length(y_interp)-start-X1(length(X1)))/fs_interp2;
recovered_fs=recovered_samples/recovered_duration;


figure(134);
f134=fft(ch1,length(X1));
fr134=recovered_fs*(0:floor(length(X1))-1)/length(X1);
plot(fr134,abs(f134(1:floor(length(X1)))));
xlabel('frequency (Hz)');
title('Ch1 recovered pre-filtering abs(fft)');

f135=f134;
```

```
fr135=fr134;


f135(round(50*recovered_duration+1))=0;
f135(round(length(X1)-(50*recovered_duration-1)))=0;

f135(round(50*recovered_duration+2))=0;
f135(round(length(X1)-(50*recovered_duration)))=0;

f135(round(50*recovered_duration+3))=0;
f135(round(length(X1)-(50*recovered_duration+1)))=0;

f135(round(50*recovered_duration-1))=0;
f135(round(length(X1)-(50*recovered_duration-3)))=0;


f135(round(50*recovered_duration))=0;
f135(round(length(X1)-(50*recovered_duration-2)))=0;


f135(round(60*recovered_duration))=0;
f135(round(length(X1)-(50*recovered_duration-2)))=0;


f135(round(100*recovered_duration+1))=0;
f135(round(length(X1)-(100*recovered_duration)))=0;

f135(round(100*recovered_duration+2))=0;
f135(round(length(X1)-(100*recovered_duration+1)))=0;


f135(round(100*recovered_duration+3))=0;
f135(round(length(X1)-(100*recovered_duration+2)))=0;



f135(round(100*recovered_duration))=0;
f135(round(length(X1)-(100*recovered_duration-1)))=0;


f135(round(150*recovered_duration+1))=0;
f135(round(length(X1)-(150*recovered_duration)))=0;

f135(round(150*recovered_duration+2))=0;
f135(round(length(X1)-(150*recovered_duration+1)))=0;
```

```
f135(round(150*recovered_duration))=0;
f135(round(length(X1)-(150*recovered_duration-1)))=0;


f135(round(200*recovered_duration+1))=0;
f135(round(length(X1)-(200*recovered_duration)))=0;

f135(round(200*recovered_duration+2))=0;
f135(round(length(X1)-(200*recovered_duration+1)))=0;

f135(round(200*recovered_duration))=0;
f135(round(length(X1)-(200*recovered_duration-1)))=0;



ch1_notched=ifft(f135,floor(length(X1)));
figure(136);
plot(fr135,abs(f135(1:floor(length(X1)))));
xlabel('frequency (Hz)');
title('Ch1 recovered & notched at 50,100,150,200Hz abs(fft)');

figure(137);
plot(X1,real(ch1_notched));
title('Ch1 recovered & notched at 50,100,150,200Hz');

figure(112);
plot(y_interp);
hold on;
plot(X1,real(ch1),'or');
title('interpolated modulated signal (blue), Ch1 recovered (red)');

figure(113);
plot(y_interp);
hold on;
plot(X1,real(ch1_notched),'or');
title('interpolated modulated signal (blue), Ch1 recovered & notched (red)');

figure(114);
plot(X1,ch1);
title('recovered Ch1');


[v,u]=cheby1(3,0.5,BW/(recovered_fs/2));
figure(31);
subplot(2,1,1);
plot(abs(u));
```

```
subplot(2,1,2);
freqz(v,u,512,recovered_fs);
grid on;
title('Ch1 LPF freq response');

ch1_LPF=filter(v,u,ch1_notched);
figure(32);
plot(real(ch1_notched));
hold on;
plot(real(ch1_LPF),'--r');
title('Ch1 (B) Vs. post-LPF (R)');

ch1_LPF_demeaned=ch1_LPF-mean(ch1_LPF);
figure(33);
plot(real(ch1_LPF_demeaned));
title('Ch1 post LPF and Demeaning');

figure(34);
f34=fft(ch1_LPF_demeaned,8192);
 fr34=recovered_fs*(0:4095)/8192;
plot(fr34,abs(f34(1:4096)));
xlabel('frequency (Hz)');
title('Filtered Ch1 abs(fft) ');

if(channels==2)
   figure(300);
   plot(y_interp);
   hold on;
   plot(X2,ch2,'or');
   hold on;
   plot(X2ref,Ch2ref,'ok');
   title('interpolated modulated signal (blue), recovered Ch2 (red)');

   figure(301);
   plot(X2,ch2);
   title('recovered Ch2');


   [v,u]=cheby1(2,3,BW/(recovered_fs/2));
   figure(302);
   subplot(2,1,1);
   plot(abs(u));
   subplot(2,1,2);
   freqz(v,u,512,recovered_fs);
   grid on;
   title('Ch2 LPF freq response');
```

```
    ch2_LPF=filter(v,u,ch2);
    figure(303);
    plot(ch2);
    hold on;
    plot(ch2_LPF,'--r');
    title('Ch2 (B) Vs. post-LPF (R)');

    ch2_LPF_demeaned=ch2_LPF-mean(ch2_LPF);
    figure(304);
    plot(ch2_LPF_demeaned);
    title('Ch2 post LPF and Demeaning');

    figure(305);
    f34=fft(ch2_LPF_demeaned,8192);
    fr34=recovered_fs*(0:4095)/8192;
    plot(fr34,abs(f34(1:4096)));
    xlabel('frequency (Hz)');
    title('Filtered Ch2 abs(fft) ');
end
```

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