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A 2.4GHz Non-Contact Biosensor System for Continuous Monitoring of Vital-Signs

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1. Introduction

In this chapter, we present a novel Doppler-based vital signs biosensor that can monitor the respiration and heartbeat rates of a person remotely without the need of any obstructions like patches, cords, etc. We will discuss the sensor operation principle and present three generations of systems that were designed to accurately extract the respiration rate and heartbeat of subjects using the Doppler radar principle. The systems have been realized using discrete custom off the shelf (COTS) parts. The first generation of the biosensor system consisted of discrete RF components system and a bulky SRS560 amplifier and filter box. Later generations of sensors consisted of custom designed printed circuits boards (PCBs) for the Doppler transceiver and for performing the analog signal processing. The data obtained using these non-contact biosensor systems was processed and logged in real-time using a LabVIEW® Graphic User Interface (GUI). Digital signal processing extracts the vital signs by filtering, auto correlating and calculating the Discrete Fourier Transform (DFT) of the waveforms. A comparison of performance among the three different generations of sensors shows that a quadrature transceiver system using autocorrelation can extract the respiration rates and heartbeat rates most accurately. Our single PCB version of the biosensor system was found to perform as well as the system using bulky components and SRS560 box. Good data accuracy has been observed on the quadrature radar sensor system with mean detection errors for respiration rate within ~1 beat/min and for heart rate within ~3 beat/min. The continuous vital signs data measured from these portable sensors can also be wirelessly transferred to healthcare professionals to make life saving decisions and diagnosis of symptoms. In the future, our vision is that a continuous log of these vital signs info can also be used to remotely monitor and gauge the recovery of patients, and even for the prevention or prediction of severe illnesses and complications.

1.1 Why non-contact vital signs monitoring?

Vital signs are measures of various physiological statistics often taken by the health professionals to assess the most basic body functions. Typical vital signs measured are: temperature, respiratory rate, heart-beat rate (i.e., pulse rate), blood pressure, and blood oxygen saturation. These numbers provide critical information about a patient's state and

healthcare professionals (especially surgeons and anesthesiologists) value the vital-signs monitoring not only during the course of surgery, but also before and after a surgery is performed (Dorin, 2007). Tremendous progresses have been made recently in developing small and wearable sensors that can monitor the vital signs; however, their usage in either hospitals or homes is still rather uncommon. One reason is that many patients are not comfortable wearing the sensors, as they have leads and are attached to the patients using patches which cause discomforts. Unless implanted, the leads and the sensors can obstruct the free movement of the patient, causing irritations or distress. Also, in situations when patients have severe burns or injuries, it can be very difficult to attach these wearable sensors to the patients. Another major disadvantage of the wearable or implantable sensors is the battery lifetime. Most wearable devices cannot operate for more than 2-3 months on a single battery for continuous and intelligent monitoring, especially if sophisticated algorithms and wireless features are added on those sensors, draining significantly more power. Continuous vital signs monitoring on patients with pre-existing complications (e.g., seizure, stroke) can ensure timely treatment, and precious lives can be saved when early warnings are triggered. The focus of this chapter is to show our development of a novel *non-contact* biosensor technology for continuous vital signs monitoring.

The principle of non-invasive microwave measurement of respiration has been described in details since 1975, while the first non-contact microwave physiological motion sensor IC was not reported until 2001 (Lin, 1992; Droitcour A.D., et al., 2002). These non-contact sensors work on the principle of Doppler shift to sense heartbeat and respiration rate by monitoring the contraction and expansion of the chest wall. However, no such biosensor product has ever been successfully commercialized for heartbeat monitoring, probably due to the practical issues caused by background clutter, phase-nulling and DC offsets. We will demonstrate in this chapter that a quadrature transceiver sensor system with arc-tangent demodulation can be very promising for implementing this kind of biosensor with robustness and reliability (Park et al., 2007). In this work, we demonstrate the design, fabrication and testing of custom wireless sensors for non-contact monitoring of vital signs. These non-contact biosensors have a range of about 1 m and can be mounted on the ceiling or unto the patient's bed in either a hospital, nursing home or at the patient's residence for non-invasive, continuous monitoring of the heartbeat and respiration rate. In the future, an alert system and RFID can be also integrated to notify medical professionals should irregularities occur. The advantages of wireless non-contact sensors include that there is no need to attach the sensors to patients for acquiring vital signs data and they can draw power from the wall socket or battery (depending on where they are used) for long-term continuous monitoring. The sensor size limitations are relaxed as they are not worn or implanted. All factors above can help to make the sensors mass producible at low cost.

1.2 Overview of vital signs using a doppler radar

The principle of non-contact microwave measurement of respiration has been described since the work of Lin *et al.* (Lin, 1992). These sensors work on the principle of Doppler-shift to sense heartbeat and respiration rates by monitoring the motion of the chest wall. We are implementing a Doppler radar motion sensing system by transmitting a high frequency continuous wave (CW) electromagnetic signal that is reflected off the subject and then demodulated in the receiver of our biosensor system. Based on the Doppler effects, we know that if a target changes position with respect to time but has no net velocity, it gives us a reflected signal whose *phase* is in proportion to the time-varying position of the target. In

our case as shown in Figure 1, a stationary person's chest has a periodic movement but no net velocity, and a CW radar system pointed to the chest as the target will receive a signal $\Phi(t)$ with its phase modulated by the time-varying chest position $x(t)$. On, multiplying the received signal with the original un-modulated signal from the same source, we can demodulate this phase information affected by the movement of the chest wall during heartbeat and respiration, and therefore heart and respiration rates can be determined. The peak-to-peak chest motion due to respiration in adults ranges from 4 mm to 12 mm (Droitcour, 2006), while the peak-to-peak motion due to the heartbeat is about 0.5 mm.

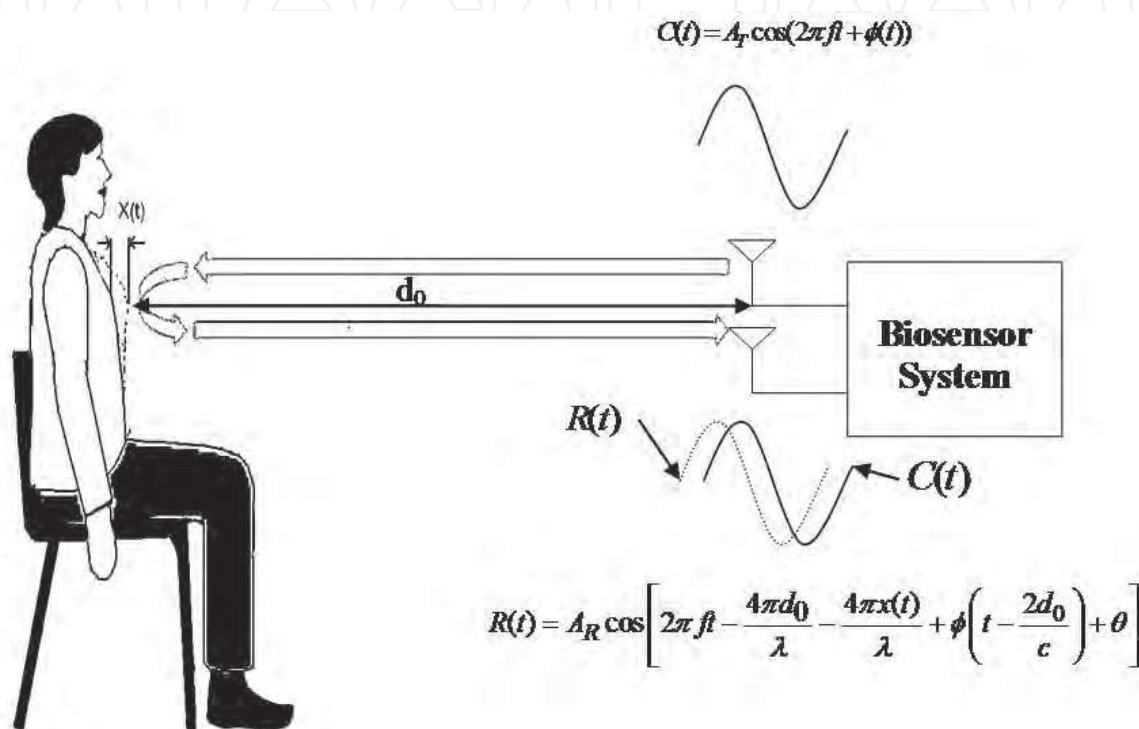


Fig. 1. Principle of operation of a Doppler-based noncontact vital-signs monitoring system

1.3 Applications of noncontact vital sign monitors

A good amount of research has been carried out recently and significant advancement has been made which have made it possible to use this Doppler technique to remotely monitor vital signs of a patient. The focus of this chapter is show how we have developed a portable non-contact biosensor technology for *continuous* vital signs monitoring. For severely burned patients and patients who suffer from serious Stevens–Johnson syndrome (SJS), it may be very difficult to attach the electrodes for vital signs monitoring; in these cases the Doppler-based non-contact vital signs monitor can be ideal to monitor the patient, reduce the risk of infection and can be safely monitored for longer duration of time. Another major advantage of our non-contact biosensor system is that it is very portable and can operate with low power, which makes it possible to monitor continuously for several weeks without having to replace batteries or power from a wall outlet. These Doppler radar based techniques can also detect *motion* and used to monitor patient movement. This can be a valuable asset to both surgeons and anesthesiologists as in some rare cases patients can start to move or have convulsions during the operating procedure which can be extremely dangerous. The Doppler radar can detect minute movements and can trigger an early warning to the

anesthesiologist/surgeon to make sure the patient is sedated. This motion sensing feature can also be used as a fall detection and prevention device (Wu et al., 2010).

Continuous monitoring of patients with preexisting conditions and complications (say, seizures, strokes, etc.) can ensure timely treatment to save precious lives as it can trigger early warnings to the caregivers and medical professionals (i.e., wireless-acute care). If continuous monitoring of vital signs is provided, a patient can be discharged early or sent to normal wards after surgical procedures, this will not only free up valuable space in the ER, but also save the patients considerable amount of money as ER care is very expensive. Other important applications for our biosensors include monitoring for sleep related disorders such as sleep apnea, sudden-infant-death (SID), etc. In addition, the vital sign sensors used in many hospitals throughout the world are not reliable as they generate a large number of *false positives*; the noncontact vital sensor will be useful and *complementary* to the traditional vital signs monitors as a perfect first-line-of-defense system to reduce the false positives of the vital signs monitoring system. This is especially true for the respiration rate monitoring. To summarize, the advantages of wireless non-contact vital sign biosensors are many as they do not have to be attached to patients for monitoring vital signs and therefore are completely non-invasive. They are very portable and can operate for a long duration, drawing power from the wall outlet or battery depending on where they are used. These biosensors can even be realized on tiny integrated circuits (ICs) and be mass produced, making them very affordable. All these factors make the non-contact vital signs monitoring very attractive for the healthcare applications.

Currently, we have tested the prototype on a limited subject pool and have gotten some good results. We have to prove the accuracy of our system for vital signs monitoring for a lot more cases, though, to show its robustness. Therefore, we need to conduct extensive trials of our system on a diverse subject pool. We have to classify the system in terms of true and false positives and negatives in a clinical environment before qualifying and commercializing this sensor in the market place. Take one case for example to illustrate the serious needs for this kind of non-contact vital signs sensor, the Texas Tech Health Science Center and the nearby University Medical Center (UMC) system have more than 35 million patient admissions a year, while the magnitude of this problem of un-monitored patients (i.e., non-ICU and non-telemetry) is large and growing and has to be properly addressed, as they consist of approximately 10-20% of overall hospital codes! We believe that the portable non-contact vital signs and motion monitor has many advantages over the current monitoring systems and therefore are extremely attractive.

1.4 Current developments in noncontact vital sign monitoring

In the late 1990s, researchers in Bell Labs worked on integrating the Doppler radar sensing function in cell phones and other portable wireless communications devices to detect the user's heartbeat and respiration rates (Li et al., 2009). Using the BiCMOS chip set developed for cellular basestation RF transceiver front end specifications, an IC for the physiological motion sensing was developed and reported in 2001 (Droitcour et al., 2002). The chipset included a low noise amplifier (LNA), a double-balanced resistive mixer, a voltage-controlled oscillator (VCO), and an active balun buffer amplifier. The IC design was optimized for low noise and high linearity for the purpose of shrinking cellular network basestation's RF front ends, but it was demonstrated to also be suitable for building noncontact Doppler radar sensors (Li et al., 2009). To simplify the design, a direct-conversion based architecture was used. One of the critical challenges involved in

integrating the microwave radar transceiver in a CMOS chip was the concern of high level of phase noise of CMOS VCO (Droitcour et al, 2004). Phase noise in the transmitted signals translates into amplitude noise on the baseband output signals in a radar system. It was shown that when measuring subjects at close distance, the range correlation effect reduces the adverse effect of phase noise significantly. Due to the range correlation effect the phase noise of the transmitted signal and the reference signal for the receiver from the same free-running VCO are correlated, making it possible to use an integrated free-running VCO. As a result, if we use the same VCO for the transmitted signal and as the LO (local oscillator) for the receive signal mixing, no phase-locked precision oscillator is needed for the non-contact sensor. This makes the fully integrated sensor chip easier. Another challenge is the detection null point. The null points occur at every quarter wavelength from the radar sensor to the subject. The detection accuracy drops significantly at these null points. To overcome this issue, quadrature receiver architecture was adopted to ensure that at least one of the two outputs is not at null (Park et al., 2007).

A. Arctangent demodulation:

It was recently demonstrated (Park et al., 2007) that arctangent demodulation in quadrature receivers achieves a high degree of accuracy in demodulation of small (heart signals) and large (walking) motion. By applying the arctangent operation to the ratio of I and Q output data, accurate phase demodulation can be achieved regardless of the target's position or motion amplitude. A particular challenge in this technique, however, is the presence of DC offset resulting from receiver imperfections and clutter reflections, in addition to DC information related to target position and associated phase. These DC components can be large compared to the ac motion-related signal, and thus cannot simply be included in digitization without adversely affecting the resolution. We have developed a method for calibrating the DC offset while preserving the DC information and capturing the motion-related signal with maximum resolution. Experimental results demonstrated that arctangent demodulation with DC offset compensation achieved a significant improvement in heart rate measurement accuracy over quadrature channel selection, with a standard deviation of less than 1 beat/min over a small number of "normal" people (i.e., 5 graduate students) for 40 to 60 sets of data is included in this chapter.

B. Double-sideband transmission/detection and frequency tuning technique:

According to the Doppler radar theory, when the displacement of physiological motion is small compared to the wavelength and the received signal is in quadrature with the reference ($\phi=90^\circ$), the small angle approximation is valid and the received baseband signal is proportional to the displacement. Therefore, one would suggest that by decreasing the wavelength, the received baseband signal strength would increase, which means smaller wavelength would be more sensitive to the small physiological motions. To verify this, a radar sensor system operating at Ka-band was developed and reported (Xiao et al., 2005). It showed that the detection range can be extended without increasing the transmitted power or antenna gain. In fact, the transmitted power can be reduced. Detection distance of greater than 2m with less than 20- μ W transmitted power as reported. That system used double-sideband transmission and detection method, which was found to achieve similar effect as the quadrature detection to avoid null-point problem when the difference in two sideband frequencies was properly adjusted. The frequency tuning technique (Xiao et al., 2006) was applied to adjust the separation of two sideband frequencies by tuning the IF LO. Although

it is indirect conversion, the double-sideband architecture is also suitable for monolithic integration since no image reject filter is needed.

C. Complex signal demodulation and random body movement cancellation:

Complex signal Demodulation consists of expressing the received I and Q signals in the time domain in the form of Bessel functions using zero, first and second order Bessel functions (Li & Lin, 2008a) as shown below:

$$I(t) = \cos\left(\frac{4\pi x_h(t)}{\lambda} + \frac{4\pi x_r(t)}{\lambda} + \phi\right) = \sum_{k=-\infty}^{\infty} \sum_{l=-\infty}^{\infty} J_k\left(\frac{4\pi m_r}{\lambda}\right) J_l\left(\frac{4\pi m_h}{\lambda}\right) \cos(k\omega_r t + l\omega_h t + \phi)$$

$$= -2[C_{10} \sin(\omega_r t) + C_{01} \sin(\omega_h t) + \dots] \cdot \sin \phi = -2[C_{20} \cos(2\omega_r t) + C_{02} \sin(2\omega_h t) + \dots] \cdot \cos \phi$$

$$Q(t) = \sin\left(\frac{4\pi x_h(t)}{\lambda} + \frac{4\pi x_r(t)}{\lambda} + \phi\right) = \sum_{k=-\infty}^{\infty} \sum_{l=-\infty}^{\infty} J_k\left(\frac{4\pi m_r}{\lambda}\right) J_l\left(\frac{4\pi m_h}{\lambda}\right) \sin(k\omega_r t + l\omega_h t + \phi)$$

$$= -2[C_{10} \sin(\omega_r t) + C_{01} \sin(\omega_h t) + \dots] \cdot \cos \phi = -2[C_{20} \cos(2\omega_r t) + C_{02} \sin(2\omega_h t) + \dots] \cdot \sin \phi$$

Then a complex FFT is used to check the spectrum of the data (Li & Lin., 2008a):

$$S(t) = I(t) + j \cdot Q(t) = \exp\left\{j\left[\frac{4\pi x_h(t)}{\lambda} + \frac{4\pi x_r(t)}{\lambda} + \phi\right]\right\}$$

$$= 2j[C_{10} \sin(\omega_r t) + C_{01} \sin(\omega_h t) + \dots] \cdot e^{j\phi} + 2[C_{20} \cos(2\omega_r t) + C_{02} \sin(2\omega_h t) + \dots] \cdot e^{j\phi}$$

Advantages:

1. The residual phase will not affect the relative strength between the odd order and the even order frequency components. The desired signal components (odd order tones) will always be present in the spectrum.
2. DC offsets exist in I/Q channels can lead to errors, but in complex demodulation they only affect the DC term. Therefore, the existence of dc offset does not affect obtaining the frequency of the desired signal components.
3. Some report that Complex FFT is computationally less intensive than regular FFT.

D. Comparison of arc tangent demodulation and complex demodulation for random Body movement cancellation:

Random Body movement is compensated by using a two sensor system placed in the front and back of the subject. The complex demodulation is not affected much by presence of baseband DC offset even with random body. Thus, the complex demodulation produces better results in the experiments carried out reported by (Li & Lin, 2008b). The critical issue of using arctangent demodulation for random body movement cancellation is how the presence of the baseband dc offset will affect the detection. This technique requires accurate calibration of the dc offset in order to properly reconstruct the angular information, which is troublesome and requires recalibration with motion of subject or changes in environment. However, the advantage of Arctangent Demodulation is the ability to eliminate the harmonics and inter-modulation interference (Li & Lin, 2008b).

E. Optimal carrier frequency:

The nonlinear property of sinusoidal transfer function not only causes the undesired effect of harmonics interference, but also causes inter-modulation between the respiration signal

and the heartbeat signal. Therefore the detected strength of a desired signal (respiration or heartbeat) is determined by both the signal itself and the other signal (i.e., heartbeat affected by the much strong respiration). In contrast to a previous belief that the detection accuracy can always be increased by increasing the carrier frequency, some argues there is an optimum choice of carrier frequency (Li & Lin 2007a). At the so-called optimum carrier frequency, the heartbeat signal component can be maximized in the premise that the harmonics interference and the inter-modulation interference are not too large to affect the detection accuracy. According to simulation and experimental demonstration, the carrier frequency was found to be optimal up to the lower region of the Ka-band to improve the detection accuracy. Since each person has different signal amplitudes of physiological motions due to different body types, the carrier frequency may be tuned to optimize the detected heartbeat signal strength. Therefore, a transceiver with a broad tuning range can be desirable to optimize the performance for different subjects under test.

F. MIMO technology:

While Single Input Multiple Output (SIMO) systems in wireless communications can provide diversity gain, array gain, and interference canceling gain, they are still providing only one source signal. In the case of Doppler radar, even with a single TX (transmitter) antenna (Zhou et al., 2006), there can be many independent signals as scatterers such as objects in subject's vicinity will scatter radio waves (acting as secondary sources), resulting in independent phase shifts. When more than one target is in view, multiple TXs and RXs (receivers) providing multiple signal copies could be used to distinguish between the different sources of Doppler motion, and isolate the desired signals (Boric-Lubecke et al, 2005; Samardzija et al., 2005). Additionally a MIMO setup was utilized to cancel random body movement and was reported in (Li & Lin., 2008b). Two Doppler transceivers at the front and back of the subject were used to correlate the body movements. By exploring the multiple antenna systems and SIMO/MIMO signal processing, it has been experimentally demonstrated that it is possible to separate physiological signals from multiple subjects (Zhou et al., 2006; Boric-Lubecke et al, 2005; Samardzija et al., 2005).

G. UWB radar approach:

Ultra-wideband (UWB) radar can also be used for implementing a non-contact physiological motion sensing. Unlike the previously described continuous-wave (CW) approach, UWB radar transmits the signal in single pulses (Staderini, 2002). By analyzing the reflected waveforms, the small movement can be extracted. The potential advantage of the UWB approach is its high sensitivity to environmental changes, making it possible to effectively remove noises caused by environmental movements and motion artifacts, etc. (Immoreev, 2007; Pisa et al., 2008; Ossberger et al., 2004). However, UWB radar requires more complicated circuits to realize desired function such as pulse shaping and delay control.

2. Theoretical background and architectural selection

2.1 Continuous wave doppler radar

Radar is the acronym for Radio Detection and Ranging. Radar is an electromagnetic system that utilizes high frequency radio waves for the detection of objects. It operates by transmitting a wave or pulses and detecting the signal reflecting back from the object. Based on correlating the reflected signal and the transmitted signal various parameters like distance of the object, velocity of the object if it is in motion, area of cross section etc may be obtained. These radar systems have several advantages over optical line-of-sight based

instruments, such as the ability to penetrate through some obstacles, operability with minimum performance degradation in rain, snow, and darkness, etc. Initially radar systems were mainly used in military applications but nowadays they find applications in areas of remote sensing for weather and exploration, air traffic control in airports, law enforcement for regulating speed on roads and highways, aircraft safety and navigation, nautical ranging and detection, space applications and recently for non-contact monitoring of vital signs (Skolnik, 1980). Radar systems can be generally divided into two types based on the signal being transmitted. They are pulsed radar systems and continuous-wave Radar systems. In a pulsed radar system, a short pulse is transmitted and there is a delay between consecutive pulses when the radar system tries to detect a reflection. In a continuous wave radar system, the radar emits a continuous wave and relative motion between the system and the object is detected either as a frequency shift or a phase shift. These radar systems cannot detect completely stationary objects as they will not cause any Doppler shifts in frequency, however they can still be very useful to detect the motion of the periodic chest wall of a stationary subject which causes a continuous phase shift depending on the position of the chest wall on which the continuous wave radar is incident. This is the main reason we find that most of the non-contact vital sign monitors employ a CW radar system to detect the respiration and heart rate of the subject. Another advantage of the CW system is that they are generally easier and cheaper to manufacture compared to pulsed radar based systems. However, the CW radar has some disadvantages too, and the major disadvantage being the coupling from the transmit chain into the receive chain either on the board itself or through the antennas which are generally closely spaced to minimize the size of the sensor system. This coupling of transmit power into the receiver can result in a time-varying or a constant DC offset, and this can limit the accuracy of detection significantly.

2.2 Principles of measurement of vital signs using a doppler radar

The Doppler shift in frequency in the general case is given as (Droitcour, 2006):

$$f_d(t) = \frac{2f}{c} v(t) = \frac{2v(t)}{\lambda} \quad (1)$$

Here $v(t)$ is the velocity of the target, λ the wavelength of the incident wave, and c is the velocity of the EM wave. However if the target is assumed to be stationary, this frequency shift can be viewed as nonlinear phase modulation as the phase signal $\Phi_r(t)$ is described as:

$$\Phi_r(t) = \frac{2f}{c} (2\pi x(t)) = \frac{4\pi x(t)}{\lambda} \quad (2)$$

Here $x(t)$ is the displacement of the chest wall. The transmitted signal from a continuous wave radar system is:

$$C(t) = A_T \cos(2\pi f t + \phi(t)) \quad (3)$$

Here f =frequency of operation and $\Phi(t)$ is the phase noise of the VCO and A_T is the amplitude of the transmitted signal. If we assume the distance from the subject to the biosensor system is d_0 and the time varying displacement of the chest is given as $x(t)$ then the distance travelled by the signal between the biosensor system and the target (chest wall of the subject under test) is given by:

$$d(t) = d_0 + x(t) \quad (4)$$

Now, the time delay between the transmitter and target is the distance travelled by the signal divided by the velocity of the wave c . However, as the chest has a time-varying displacement, the distance travelled by the signal at the time of reflection is given as:

$$d(t_{rf}) = d\left(t - \frac{d(t)}{c}\right) \quad (5)$$

Therefore the time delay by the time the signal is received is given as:

$$t_d = \frac{2d\left(t - \frac{d(t)}{c}\right)}{c} = \frac{2\left(d_0 + x\left(t - \frac{d(t)}{c}\right)\right)}{c} \quad (6)$$

The signal at the receiver $R(t)$ is a time-delayed version of the transmitted signal given by equation (3). We get the received signal to be:

$$R(t) = A_R \cos\left[2\pi f\left(t - t_d\right) + \phi\left(t - t_d\right) + \theta\right] \quad (7)$$

Here A_R is the amplitude of the received signal. Substituting for t_d from Eq.(6) in Eq. (7):

$$R(t) = A_R \cos\left[2\pi f\left(t - \frac{2\left(d_0 + x\left(t - \frac{d(t)}{c}\right)\right)}{c}\right) + \phi\left(t - \frac{2\left(d_0 + x\left(t - \frac{d(t)}{c}\right)\right)}{c}\right) + \theta\right] \quad (8)$$

$$R(t) = A_R \cos\left[2\pi ft - \frac{4\pi d_0}{\lambda} - \frac{4\pi x\left(t - \frac{d(t)}{c}\right)}{\lambda} + \phi\left(t - \frac{2d_0}{c} - \frac{2x\left(t - \frac{d(t)}{c}\right)}{c}\right) + \theta\right] \quad (9)$$

Here, the wavelength $\lambda = c/f$, and we know that the period of the chest movement for both heartbeat and respiration has a time period $T \gg d_0 / c$, from this we can neglect the

$4\pi x\left(\frac{d(t)}{c}\right)$ term in Eq. (9) and similarly in the phase noise term we can neglect

$\phi\left(2x\left(t - \frac{d(t)}{c}\right)/c\right)$ as the displacement of the chest movement generally in the order of 1 cm

is much smaller compared to the distance between the subject and the biosensor system which is generally 50 cm to 2m. On omitting these terms in Eq. (9) for reasons described above, the received signal becomes:

$$R(t) = A_R \cos\left[2\pi ft - \frac{4\pi d_0}{\lambda} - \frac{4\pi x(t)}{\lambda} + \phi\left(t - \frac{2d_0}{c}\right) + \theta\right] \quad (10)$$

From Eq. (10), we can see that the received signal $R(t)$ resembles a time delayed version of the transmitted signal with *phase modulation*. The time delay is determined mainly by the distance d_0 between the biosensor system and the subject. The periodic motion of the target $x(t)$ is superimposed as phase modulation. In order to determine the heart rate and respiration rate we need to demodulate this phase modulation to correlate the motion of the chest wall with actual heart rate and respiration rate.

From Eq. (10) we can see the wavelength has to be greater than at least twice the peak-to-peak motion $x(t)$ of the chest wall. Otherwise we may get aliasing in the demodulated signal caused due to the under sampling of the phase modulated information by the sampling wave. So the lower limit on the possible choice of frequencies exists. The higher limit on the possible choices of operating frequency is given by the wavelength where the small signal approximation can no longer be applied which would complicate the analysis of the system.

2.3 Transceiver architectures and design

In the present section, we will discuss the choice of transceiver architectures and try to understand the advantages and disadvantages of each option in our attempt to determine which suits our application the best. As we have seen from the mathematical analysis above, the noncontact biosensor system needs to transmit CW at a fixed frequency. It needs to demodulate the phase modulated motion of the chest wall to determine the heart rate and the respiration rate. Transmitting CW at a fixed frequency can be easily accomplished by using a VCO tuned to the desired frequency. As Eq. (10) implies, the phase noise of the oscillator is an important parameter. If the VCO output is also used to drive the receive Local Oscillator (LO) inputs of the receive mixers, this will eliminate the need for a Phase-Locked-Loop (PLL) and also has an additional benefit of relaxing the phase noise requirements of the VCO by the phenomenon of range correlation. Depending on the range being targeted, a power amplifier can be added to boost output power. In order to obtain the vital sign information we need to down convert the received RF signals into baseband. This can be easily accomplished by mixing the received signal with a signal at the same carrier frequency, resulting in a conversion directly to baseband signals. This type of receiver is known either as a direct conversion or a homodyne receiver. A heterodyne receiver instead mixes the received signal with a LO signal at a different frequency, so the information is modulated on a non-zero intermediate frequency (IF) rather than being converted directly to baseband. This following section introduces the heterodyne and direct-conversion architectures, describes why the direct-conversion architecture is chosen for Doppler radar cardio-respiratory monitoring, and introduces the theory for Doppler radar monitoring with a CW system and a direct-conversion receiver.

2.3.1 Heterodyne transceiver architecture

In a heterodyne receiver, the input RF signal is mixed with a LO signal generated at a different frequency to the RF input signal, resulting in a signal at IF. This signal is filtered with a fixed bandpass filter and then amplified by a tuned IF stage amplifier and then demodulated directly or mixed down to baseband before demodulation. The receiver's basic architecture is shown in below. One major drawback with the heterodyne architecture is that it requires more circuits and passives as two-stage conversions are performed.

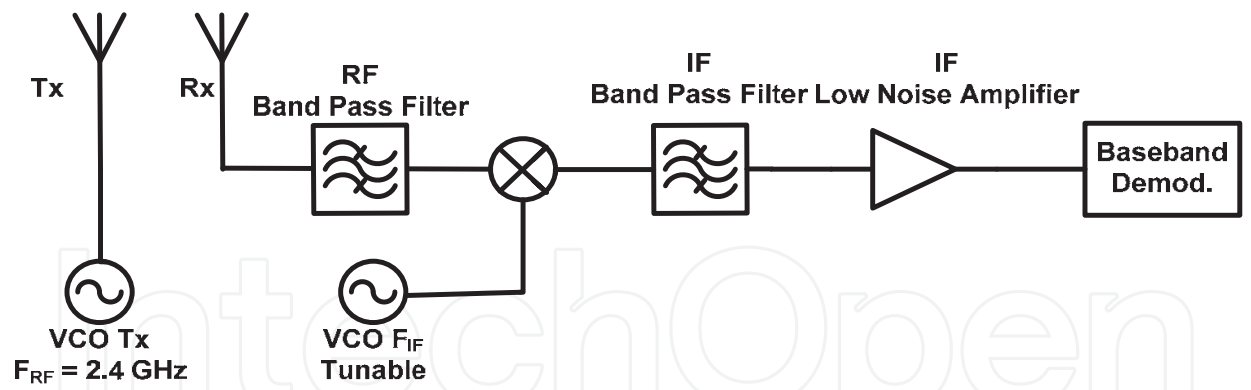


Fig. 2. Radar architecture using a heterodyne transceiver (single-channel)

2.3.2 Direct conversion or homodyne transceiver architecture

In a homodyne receiver, as shown in Figure 3, the received signal is mixed with a LO frequency at the same frequency as its carrier, converting the signal to baseband. The baseband signal is then filtered using a bandpass filter and then amplified by a baseband amplifier to the appropriate signal level for the digitizer (ADC) and demodulator. One major disadvantages of using a homodyne receiver for vital signs monitoring is the amount of DC offset introduced by the system; since the vital signs data is at very low frequency (0.3 to 3Hz), it can be difficult to demodulate the data in the presence of high DC offset so close to the desired frequency band. The DC offsets are caused by system imperfections like LO feed through, self mixing or by reflections from clutter or other nearby objects. The DC offsets can be as high as 200mV (Park et al., 2007) compared to the 1-10 mV respiration signals and the 10-100 μ V heartbeat signals. Filtering this DC offset is essential to prevent saturation of ADC.

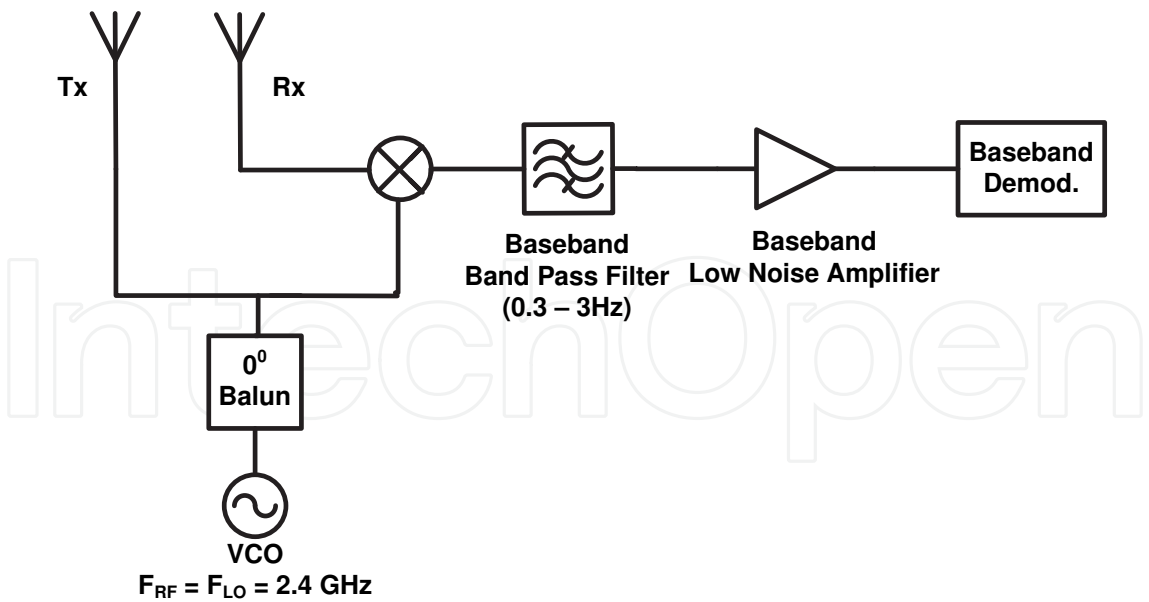


Fig. 3. Architecture of a Homodyne Transceiver

2.4 Transceiver architecture selection and design for vital signs monitoring

The direct conversion (homodyne) receiver for non-contact monitoring system has several advantages over a heterodyne receiver for applications that require continuous vital signs

monitoring and therefore was chosen as the preferred architecture for our design. Some of the major advantages are:

- a. *Range Correlation:* Range correlation is the process by which the phase noise of the transmitted VCO is correlated to the received signal cancelling out some of the phase noise of the received signal (Droitcour et al., 2003). This is an extremely valuable property of the direct conversion receivers as this can greatly relax the phase noise spec on the VCO. This is conceptually shown in the figure below.

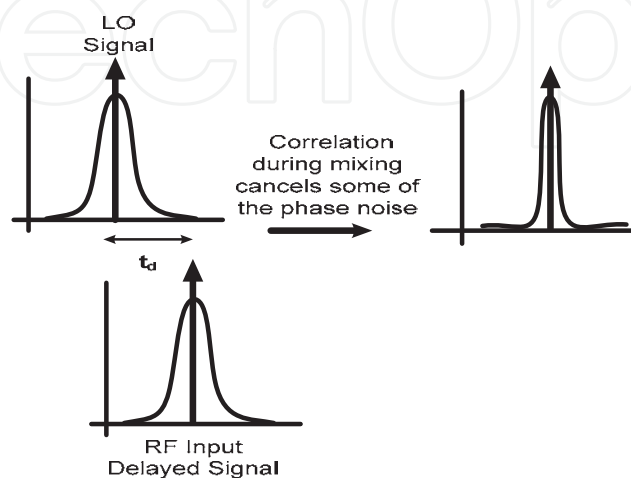


Fig. 4. Illustration of Range Correlation

- b. *Simple Architecture and Design:* The direct conversion receiver is a simpler architecture to design as it is less susceptible to image and therefore does not require a RF bandpass image filter in the front end. It also uses the same VCO for both transmission and LO generation so it does not need a separate tuning stage for the baseband LO.
- c. *Low Power Consumption:* The direct conversion receiver requires fewer components and therefore consumes lower power than a heterodyne receiver. During the operation as a portable non-contact vital sign sensor if we are operating these sensors from battery powered sources, we want to keep the power consumption to a minimum.

2.5 Single channel vs. quadrature (I-Q) radar systems

The Doppler radar transceivers can be built based on a single channel design or a quadrature design. In the system being reported in the current thesis, we built a single channel system initially and then switched to quadrature transceivers mainly to overcome the problem present in all Doppler radar transceivers: the phase-demodulation null points. These null points depend on the phase relationship between the LO and the received signal. This problem can be mitigated by using a quadrature receiver, which provides two receiver chains with a 90° difference in their LO phases.

We know from Eq 10 that the received signal which is the RF input to the mixer is:

$$R(t) = A_R \cos \left[2\pi f t - \frac{4\pi d_o}{\lambda} - \frac{4\pi x(t)}{\lambda} - \frac{4\pi y(t)}{\lambda} + \phi \left(t - \frac{2d_o}{c} \right) + \theta \right] \quad (11)$$

Here, $x(t)$ is the signal due to respiration and $y(t)$ is the signal due to the heart beat. Also the transmitted signal is used for driving the LO of the mixer, so the LO input to the mixer is:

$$C(t) = A_T \cos(2\pi ft + \phi(t))$$

On mixing these signals at the mixer and after filtering to obtain the desired low frequency components we get (Park et al., 2006):

$$D(t) = \cos \left[\theta + \frac{4\pi x(t)}{\lambda} + \Delta\phi(t) \right] \quad (12)$$

Here, $\Delta\phi(t)$ is the residual phase noise in the baseband demodulated signal (Park et al., 2006):

$$\Delta\phi(t) = \left[\phi(t) - \phi \left(t - \frac{2d_o}{c} \right) \right] \quad (13)$$

Also, θ is the constant phase shift dependant on the nominal distance to the target d_o :

$$\theta = \left[\frac{4\pi d_o}{\lambda} + \theta_0 \right] \quad (14)$$

The null and optimum demodulation points are the two extreme cases for the output signal given as a function of θ from Eq. (12). The optimum detection point occurs when θ is an odd multiple of $(\pi/2)$, In this case applying the small signal approximation as the cardiopulmonary signal information $x(t) \ll (\lambda/4)$ we get:

$$D(t) = \left[\frac{4\pi x(t)}{\lambda} + \frac{4\pi y(t)}{\lambda} + \Delta\phi(t) \right] \quad (15)$$

Now, assuming the displacements associated with respiration and heart rate can be modeled as sinusoids as a first order approximation, we have (Park et al., 2006):

$$D(t) \approx [R \sin 2\pi f_r(t) + H \sin 2\pi f_h(t) + \Delta\phi(t)] \quad (16)$$

Here, $R \gg H$ and $f_h \gg f_r$ and the demodulated output signal is linearly proportional to the cardio pulmonary information signals. For the optimum case of demodulation, we can easily demodulate the vital signs data after suitable amplification and filtering.

In the case of the null detection point, when θ is an odd multiple of π , we have

$$D(t) \approx 1 - [R \sin 2\pi f_r(t) + H \sin 2\pi f_h(t) + \Delta\phi(t)]^2 \quad (17)$$

If we assume the residual phase noise $\Delta\phi(t)$ is much smaller than the cardiopulmonary signals, we can neglect that term in Eq. 17. And expanding square terms we get:

$$D(t) \approx 1 - \left[R^2 \sin^2 2\pi f_r(t) + H^2 \sin^2 2\pi f_h(t) + 2HR \sin 2\pi f_r(t) \cdot \sin 2\pi f_h(t) \right] \quad (18)$$

Simplifying further we get:

$$D(t) \approx 1 - \frac{1}{2} \left[(R^2 + H^2) - R^2 \sin^2 4\pi f_r(t) - H^2 \sin^2 4\pi f_h(t) - 2HR(\cos 2\pi(f_h + f_r)t - \cos 2\pi(f_h - f_r)t) \right] \quad (19)$$

We can see that in the case of null demodulation point, we get a signal that has several distortion terms and the output signal is proportional to the square of the cardiopulmonary informational signals causing decreased sensitivity. Also, the only term that passes the low pass filter is the second order respiration term, along with the DC offsets associated with the output signal. It should be noted that the term being produced has twice the frequency of the true respiration signal. In addition, the respiration signals are much larger in magnitude than that of the heart rate signals, and at these null points there is a mutual coupling of the true heart and the respiration signals (which has a higher magnitude than the heart signal), making accurate demodulation of the heart rate signal very difficult at these null points. In our current system with an operating frequency of 2.4GHz, we have a λ of 12.5 cm and we know from Eq. (14) this means we have a null point every 3.125cm. Since we cannot make such fine adjustments in the positioning of the vital signs monitoring sensor system, we need to implement a quadrature transceiver system that drives the in-phase and quadrature channels with a 90° phase shift such that when one channel is at a null point we have the other channel at an optimum point for demodulation. By either selecting the output closest to the optimal phase demodulation point or by intelligently combining the two outputs, phase-demodulation null points can be avoided and accurate results for both heart rate and respiration rate can be obtained. Another reason for implementing a quadrature receiver is that in direct conversion receivers, image rejection cannot be done with filtering, because the image signal and desired signal are in the same frequency space. So by using quadrature architecture we can convert each of the in-phase and quadrature frequency components individually to baseband so that the image signal can be rejected (depending on how good the I/Q matching is) to minimize its effect on the desired signal.

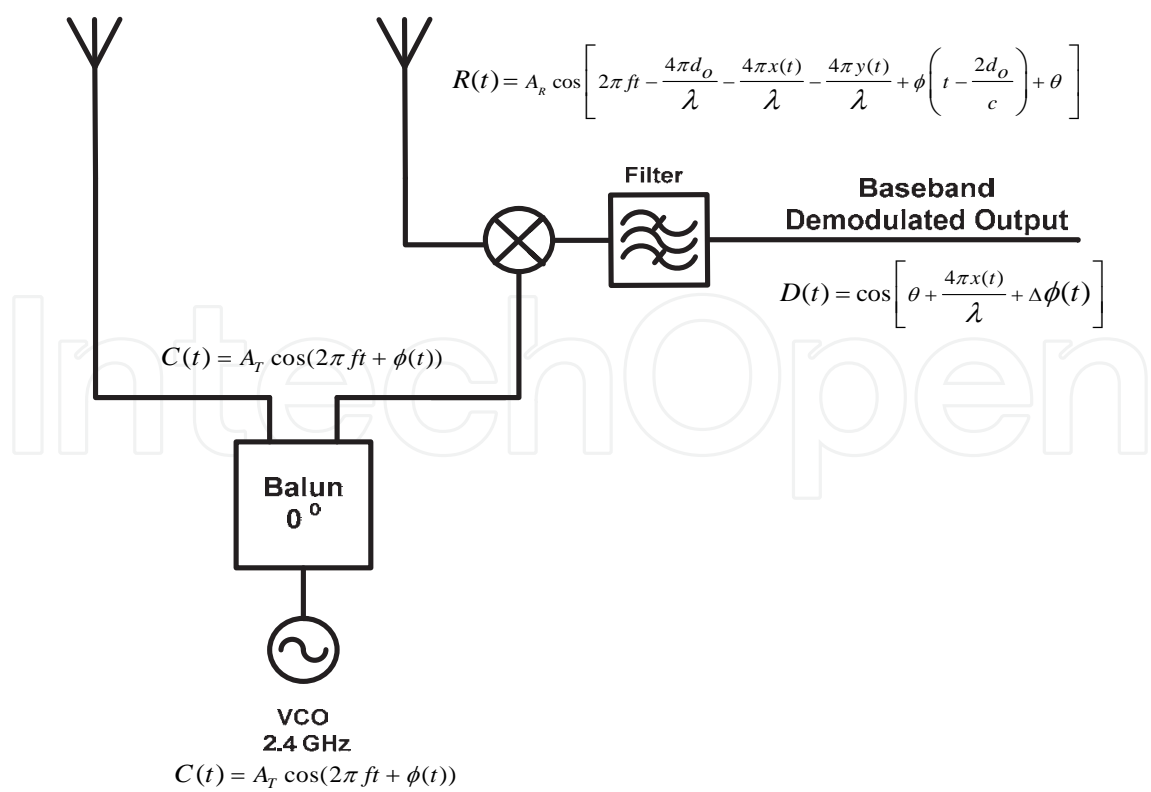


Fig. 5. System Diagram showing Demodulation with equations

2.6 Arctangent demodulation

As discussed above, the performance of single-channel Doppler Radar transceiver is known to be sensitive to the position of the target. For the worst case, the target can be at a “null” position that will produce virtually no signal due to its distance away from the radar at integer multiples of the quarter of wavelength (λ). Therefore, to avoid “optimum” and “null” extremities in the demodulated signal, we adopted a Quadrature Doppler Radar receiver system here. However, these in-phase and quadrature outputs need to be combined effectively to get accurate detection of the vital signs data. Arctangent combination of the I and Q signals has been shown to give highly accurate results (Park et al, 2007). We have also used this method to combine the quadrature data for demodulation in the present thesis work. In a quadrature system the two orthogonal baseband outputs are (Droitcour 2006):

$$I_B(t) = \cos \left[\theta + \frac{4\pi x(t)}{\lambda} + \frac{4\pi y(t)}{\lambda} + \Delta\phi(t) \right] \quad (20)$$

$$Q_B(t) = \sin \left[\theta + \frac{4\pi x(t)}{\lambda} + \frac{4\pi y(t)}{\lambda} + \Delta\phi(t) \right] \quad (21)$$

Here, θ is the constant phase shift dependent on the nominal distance to the target, $x(t)$ is the heart motion signal, $y(t)$ is the respiration motion signal, and

$$\Delta\phi(t) = \phi(t) - \phi \left(t - \frac{2d_0}{c} \right) \quad (22)$$

Here, $\phi(t)$ is the residual VCO phase noise translated in baseband, which depends on the nominal target distance. One can combine the I and Q channels in Eqs. (3) and (4) by using the arctangent function (Park et al., 2007). This gives us a resultant baseband signal:

$$\Phi_r(t) = \arctan \left(\frac{Q_B(t)}{I_B(t)} \right) \quad (23)$$

$$\Phi_r(t) = \arctan \left(\frac{\sin \left[\theta + \frac{4\pi x(t)}{\lambda} + \frac{4\pi y(t)}{\lambda} + \Delta\phi(t) \right]}{\cos \left[\theta + \frac{4\pi x(t)}{\lambda} + \frac{4\pi y(t)}{\lambda} + \Delta\phi(t) \right]} \right) \quad (24)$$

$$\Phi_r(t) = \left[\theta + \frac{4\pi x(t)}{\lambda} + \frac{4\pi y(t)}{\lambda} + \Delta\phi(t) \right] \quad (25)$$

A real system has several non-idealities like amplitude and phase mismatches and other hardware imperfections. There can be a significant DC offset term introduced by hardware imperfections and reflections from background clutter, etc. (Park et al, 2007) Due to these non-idealities, the demodulated signal using the arctangent function above changes to:

$$\Phi_r'(t) = \arctan \left(\frac{A_Q + A_E \sin \left[\theta + \frac{4\pi x(t)}{\lambda} + \frac{4\pi y(t)}{\lambda} + \Delta\phi(t) + \phi_E \right]}{A_I + \cos \left[\theta + \frac{4\pi x(t)}{\lambda} + \frac{4\pi y(t)}{\lambda} + \Delta\phi(t) \right]} \right) \quad (26)$$

Here, A_Q and A_I denote the dc offsets of Q and I channel, respectively, and A_E and ϕ_E are the amplitude error and phase error, respectively.

2.6.1 Comparison of arctangent demodulation and complex demodulation

We know from the above section that the arctangent combination gives us the resultant phase modulated vital signs information as shown in Eq. 26. Here, A_Q and A_I denote the dc offsets of I and Q channel respectively and are generally the most important parameters in determining the accuracy of the extracted heartbeat and respiration rate information. However, these DC offsets comprise of two components, one component is the “desirable DC offset” component that consists of some vital signs information and the other DC offset “undesirable DC offset” that consists of the DC offset caused by electronic components and caused by reflections from stationary or other objects in the vicinity of the subject. If this DC offset is improperly calibrated in the baseband, it leads to a shifted trajectory. When the trajectory is shifted the vital signs data is still present and can be extracted, but the difference is a changed harmonic level in the baseband spectrum. The arctangent method is reported to be more sensitive to these DC offsets especially in the presence of random body movements (Li & Lin., 2008b). Calibration of the DC offset can be a difficult problem because the DC offset is a dynamic variable and changes with environment, position of the subject, etc. To overcome the problems of DC offset calibration, we can also use a complex signal demodulation method to combine the I & Q channel data as described before.

The baseband signals can be expressed using Bessel's Functions as:

$$I(t) = \cos \left(\frac{4\pi x_h(t)}{\lambda} + \frac{4\pi x_r(t)}{\lambda} + \phi \right) = \sum_{k=-\infty}^{\infty} \sum_{l=-\infty}^{\infty} J_k \left(\frac{4\pi m_r}{\lambda} \right) J_l \left(\frac{4\pi m_h}{\lambda} \right) \cos(k\omega_r t + l\omega_h t + \phi) \quad (27)$$

$$= -2[C_{10} \sin(\omega_r t) + C_{01} \sin(\omega_h t) + \dots] \cdot \sin \phi = -2[C_{20} \cos(2\omega_r t) + C_{02} \sin(2\omega_h t) + \dots] \cdot \cos \phi$$

$$Q(t) = \sin \left(\frac{4\pi x_h(t)}{\lambda} + \frac{4\pi x_r(t)}{\lambda} + \phi \right) = \sum_{k=-\infty}^{\infty} \sum_{l=-\infty}^{\infty} J_k \left(\frac{4\pi m_r}{\lambda} \right) J_l \left(\frac{4\pi m_h}{\lambda} \right) \sin(k\omega_r t + l\omega_h t + \phi) \quad (28)$$

$$= -2[C_{10} \sin(\omega_r t) + C_{01} \sin(\omega_h t) + \dots] \cdot \cos \phi = -2[C_{20} \cos(2\omega_r t) + C_{02} \sin(2\omega_h t) + \dots] \cdot \sin \phi$$

, where $C_{ij} = J_i \left(\frac{4\pi m_r}{\lambda} \right) \cdot J_j \left(\frac{4\pi m_h}{\lambda} \right)$ determines the amplitude of each frequency component.

Hence, the DC components are given by:

$$DC_I = J_0 \left(\frac{4\pi m_r}{\lambda} \right) \cdot J_0 \left(\frac{4\pi m_h}{\lambda} \right) \cdot \cos \phi \quad (29)$$

$$DC_Q = J_0 \left(\frac{4\pi m_r}{\lambda} \right) \cdot J_0 \left(\frac{4\pi m_h}{\lambda} \right) \cdot \sin \phi \quad (30)$$

Then a complex FFT is used to combine the I & Q channels in the baseband (Li & Lin., 2008a):

$$S(t) = I(t) + j \cdot Q(t) = \exp \left\{ j \left[\frac{4\pi x_h(t)}{\lambda} + \frac{4\pi x_r(t)}{\lambda} + \phi \right] \right\} \quad (31)$$

$$= DC_{IQ} + 2j[C_{10} \sin(\omega_r t) + C_{01} \sin(\omega_h t) + \dots] \cdot e^{j\phi} + 2[C_{20} \cos(2\omega_r t) + C_{02} \sin(2\omega_h t) + \dots] \cdot e^{j\phi}$$

Since $e^{j\phi}$ has a constant envelope of one, the effect of ϕ on signal amplitude can be eliminated. Applying the complex Fourier transform to the signal for spectral analysis, the residual phase will not affect the relative strength between the odd order and the even order frequency components. The desired signal components which are the odd order tones will always be present in the spectrum (Li & Lin., 2008b). DC offsets that exist in the I/Q channels and may lead to the error of measured vital signs, only affect the dc term of $S(t)$ in Eq. (31). This DC offset can be extracted and safely removed by averaging the baseband signals and thus the existence of dc offset does not affect obtaining the frequency of the desired signal components making the complex signal demodulation more immune to DC offset mismatches. However, the complex signal demodulation method is affected by the even order harmonics that are still present in the baseband spectrum.

2.7 Summary on the architectural selection and theoretical background

A continuous-wave direct-conversion radar transceiver is the fundamental component of our non-contact vital signs monitoring system. Direct-conversion receiver architecture was chosen due to its simplicity, low power consumption, good performance, and lower fabrication cost compared to a heterodyne receiver. A quadrature receiver was used to avoid null points and it can provide us position-insensitive accurate results as null points can be difficult to avoid due to variations in the relative position of the subject (they occur every 3.125 cm at our operating frequency). Quadrature receivers consume more power and need more components making them more expensive than single channel systems.

3. Measured vital signs results and data analysis

3.1 Introduction

Our noncontact vital signs monitoring system consisted of a Doppler radar transceiver transmitting a continuous wave using a VCO, a single channel or quadrature receiver, custom analog baseband signal conditioning hardware and a custom GUI developed in Labview that performs digital signal processing, extraction of vital signs and logging of the data. For the purposes of calibration and comparison, a piezoelectric pulse transducer sensor was worn on a subject's hand and the signals from the transducer were used to compute the reference heartbeat signal. The in-phase (I), quadrature (Q) and reference channels were digitized simultaneously using a NI USB6000 ADC connected to a PC. The piezoelectric sensor was used to find the reference heartbeat rate which was compared with the rates computed by the Doppler sensor system using the Labview or Matlab software.

3.2 Experimental setup

Our experimental setup consisted of the Doppler noncontact radar transceiver, an analog signal processing system and a NI ADC connected to a PC running Labview. The subject

was seated in a chair in an upright position and the piezoelectric pulse transducer sensor UFI 1010 was affixed to the subject's right hand index finger. The subject was then asked to sit straight and still in a chair for the measurements, and the antenna placement was adjusted to be approximately at the same level with the center of the subject's chest. The primary objective of this data collection was to determine whether the Doppler noncontact vital signs monitor can accurately extract the heartbeat and respiration rates of healthy, relaxed subjects in a well controlled environment with no background motion or presence of other humans in the vicinity of the Doppler radar system. During the course of the data collection, the subject was asked to remain still and to refrain from making any involuntary motions like scratching, talking or other movements for the complete duration of each measurement. The measurement setup is shown in the figure below.

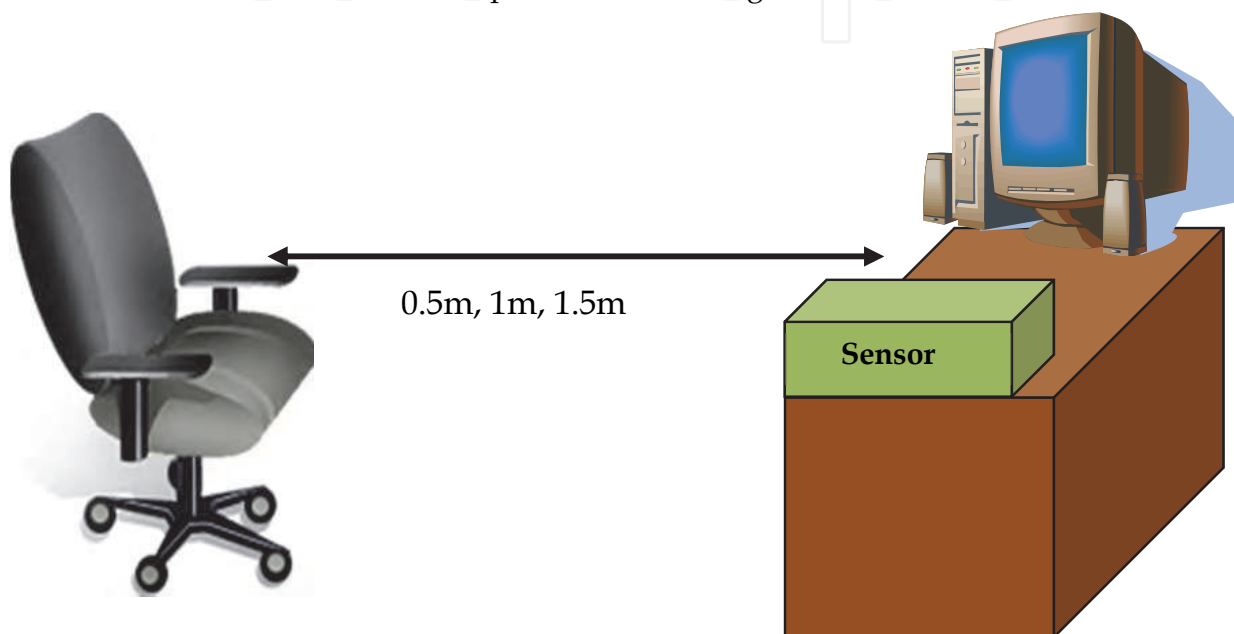


Fig. 6. Diagram showing the noncontact vital signs sensor measurement setup

3.3 Results and data analysis

In the present section we discuss the results of the data collected using our first, second and third generation systems and try to contrast and compare the differences in performance. For each reading that was collected, we compute the mean error, standard deviation and median. The mean error over several readings was computed and used to determine the accuracy of the system. As these Doppler radar based sensor systems are very sensitive to noise, background movement, stray reflections from clutters, etc., averaging the extracted respiration and heart rates becomes necessary. Momentary readings might be distorted due to sudden involuntary motion of the subject like sneezing, coughing etc. Also, another figure of merit for the sensor system is the standard deviation of the error. We can use the standard deviation as an estimate of the worst case accuracy range within which our system extracts the vital signs rates in this controlled environment.

3.3.1 First generation sensor system

This system was designed to initially show the proof of concept of the ability of extraction of vital signs using a Doppler transceiver. This system was a single channel system as shown

in figure below. This system comprised of discrete SMA terminated Mini-circuits modules assembled together using SMA cables. The Analog Signal Processing (ASP) was done initially using a SRS 560 Preamplifier and Filter block. However, since the SRS560 block is bulky and expensive, we designed a custom ASP board. We compared the performance of both systems to verify that the ASP board performs as well as the SRS560, so we have replaced it with the custom ASP board. In this section we present the data collected by the SRS560 block and the ASP board, where the data was logged using the NI Data Logger that comes bundled with the NI USB ADC and logs the sampled voltage readings from each channel into a text file. The text file was analyzed using a MATLAB script and the respiration rate and heartbeat rate were extracted using only the autocorrelation function. The extracted heartbeat was compared to the piezoelectric reference from the finger pulse transducer sensor. From the data we can clearly see that the ASP board performs as well as or sometimes appears better than the SRS560 block. The mean and the standard deviation of the error of the extracted respiration rate of the ASP board are comparable to the mean and standard deviation of the error from the SRS560 block. We can conclude that the ASP board performs well and can be used as a replacement for the SRS560 block.

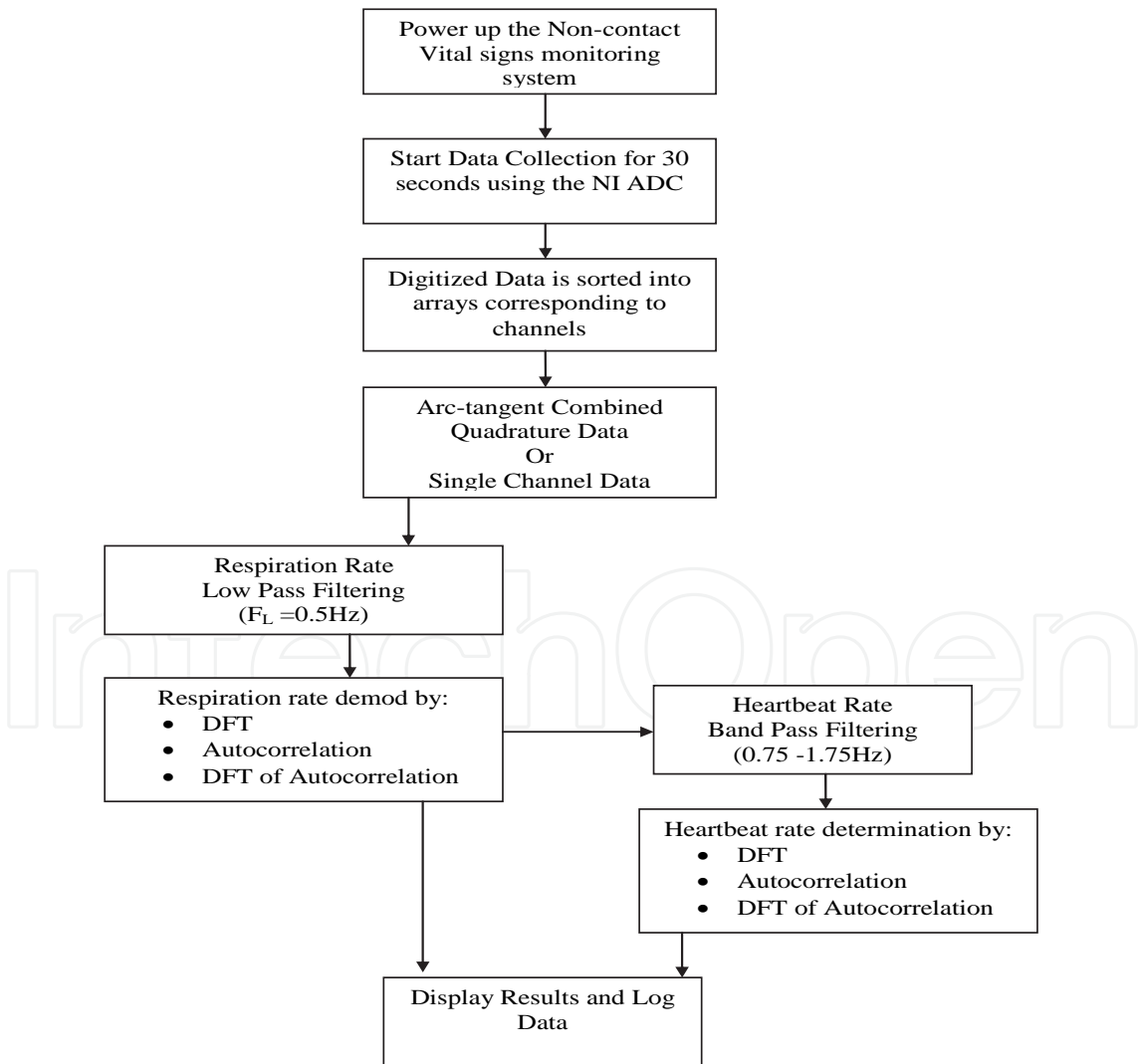


Fig. 7. Flowchart showing our sensor measurement sequence

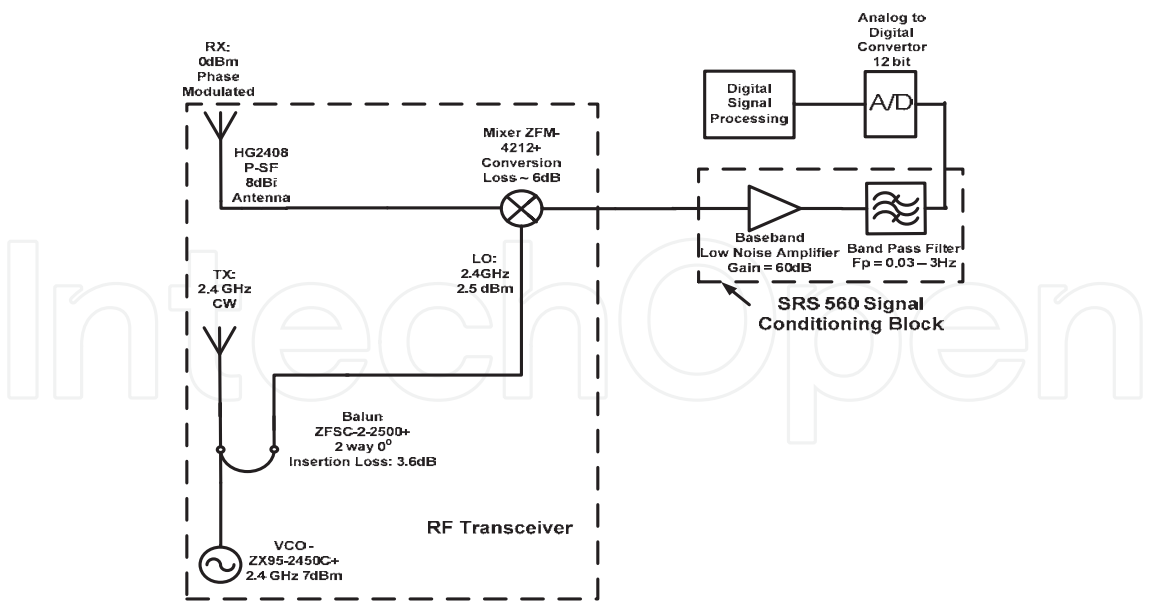


Fig. 8. First Generation noncontact vital signs system developed at Texas Tech U.

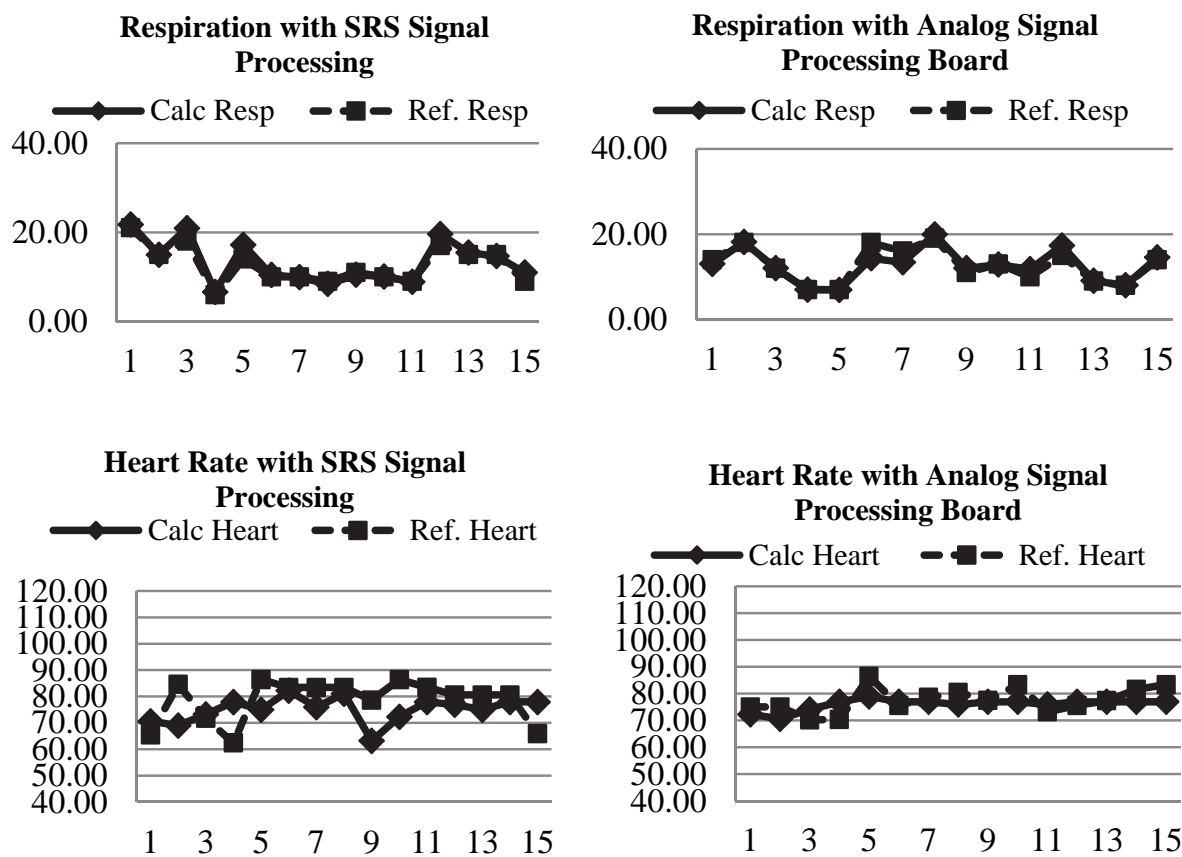


Fig. 9. (a). Respiration Data showing how well the system tracks the reference with the SRS560 Block (left); ASP board (right), (b). Heartbeat data showing how well the system tracks the reference with the SRS560 block (left); with our custom ASP board (right)

	SRS 560	ASP Board
Respiration (beat/min)		
Mean Error	0.76	-0.03
STDEV of Error	1.28	1.51
Heartbeat (beat/min)		
Mean Error	-3.38	-1.56
STDEV of Error	9.25	4.16

Table 1. Comparison of measured vital signs data using SRS560 vs. custom ASP board

3.3.2 Second generation sensor system

The second generation system is a Quadrature Doppler transceiver to get position-insensitive and accurate results compared to the first generation single channel system. It comprised of the first RF PCB that was designed to eliminate the bulkier SMA modules. We designed this system using our custom designed Analog PCB that comprises of Sallen-Key Active filters and an Op-amp based baseband Low Noise Amplifier. The system block diagram, the lab setup, the analog processing/filtering design, and pictures of two discrete RF and Analog PCBs are shown in the figures below.

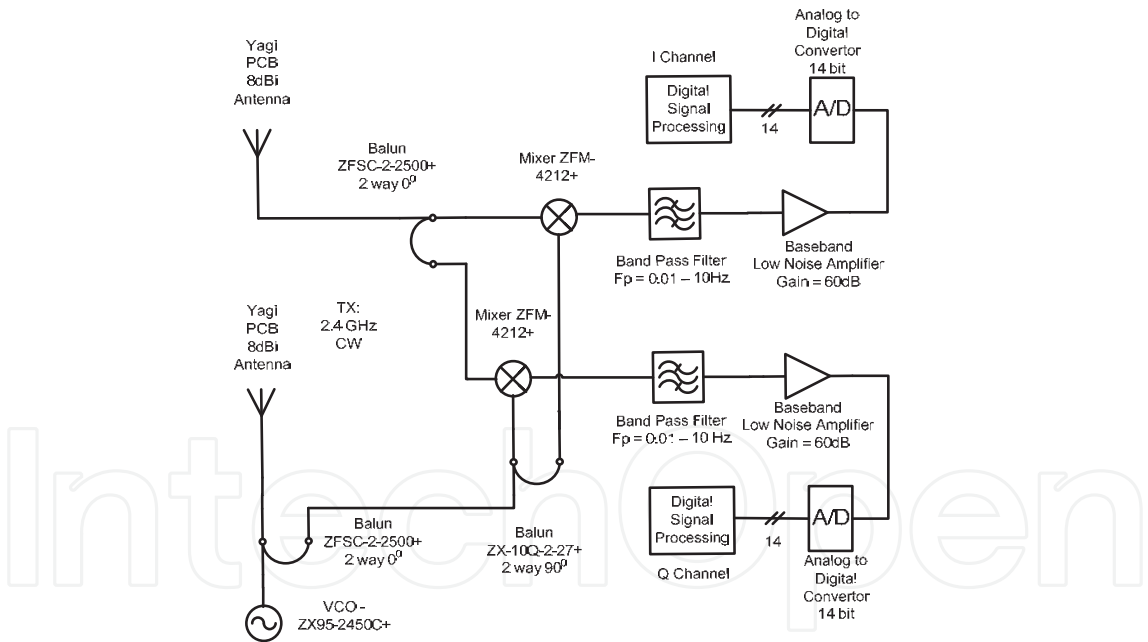


Fig. 10. Second generation sensor system with discrete components developed in Texas Tech



Fig. 11. A picture of the 2nd generation sensor with discrete components (PC not shown)

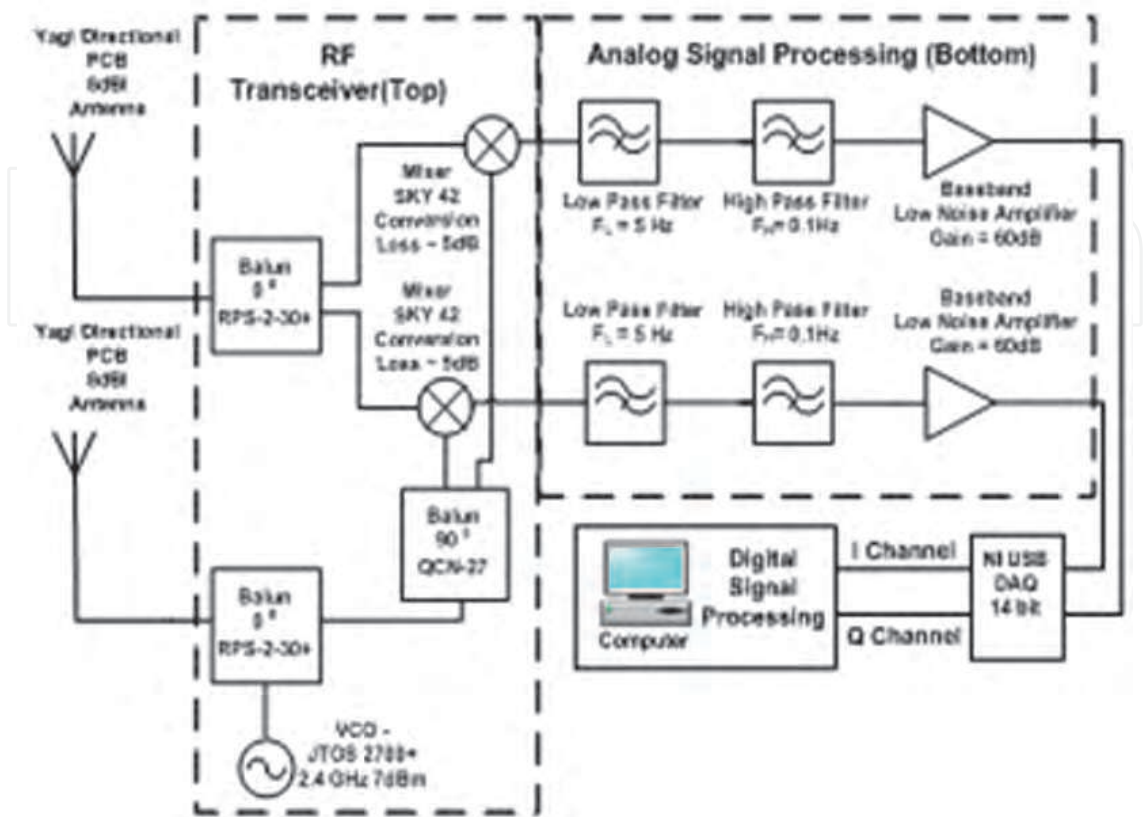


Fig. 12. Second generation sensor system with RF and Analog PCBs

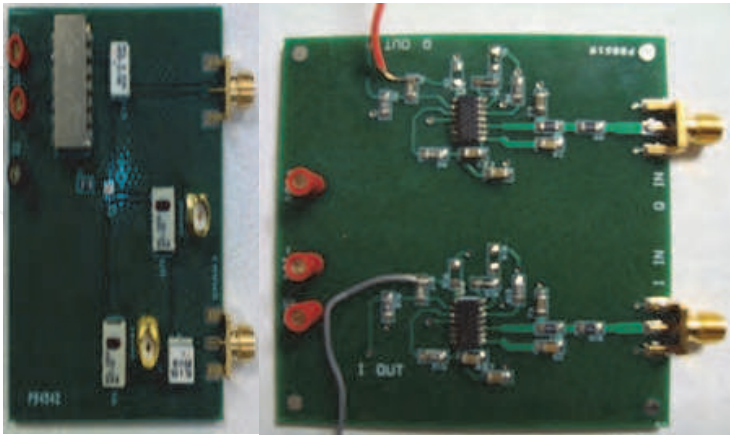


Fig. 13. Second Generation PCB system for our sensor: RF PCB (left); Analog PCB (right)

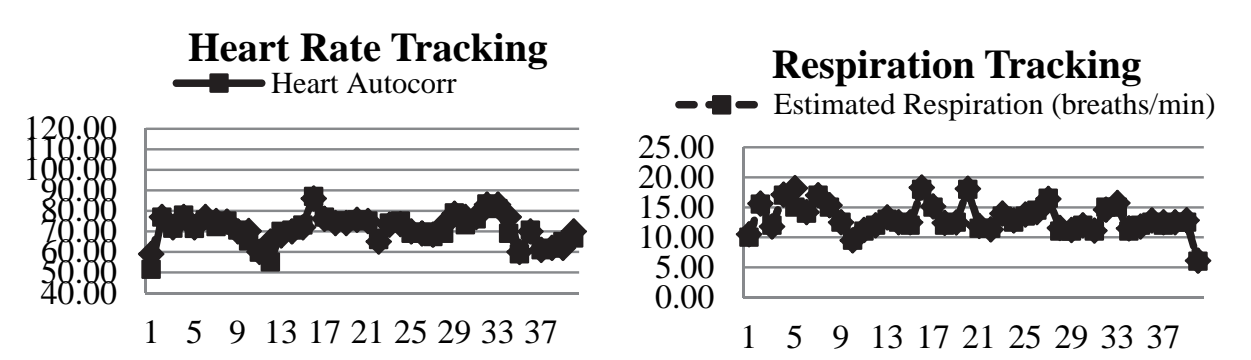


Fig. 14. 2nd Gen. sensor data tracks references for heartbeat (left) and respiration rate (right)

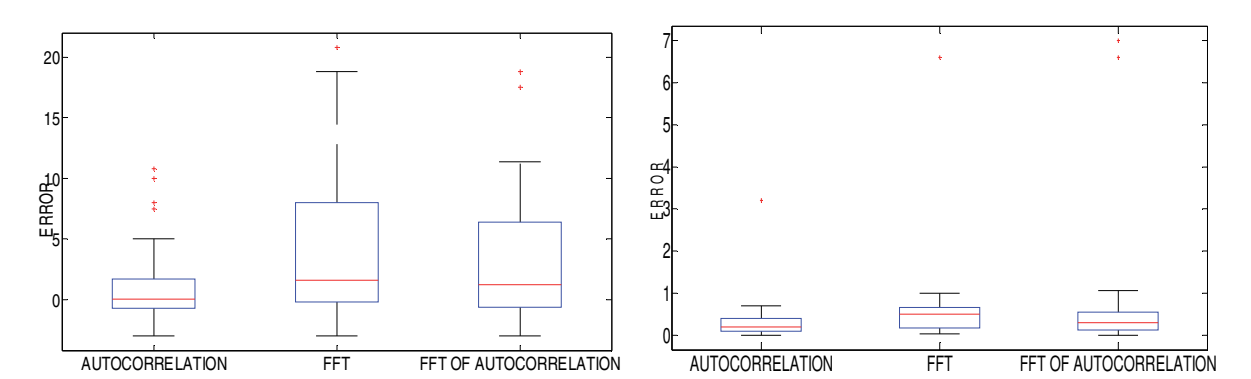


Fig. 15. Box plots of errors for 2nd Gen. sensor of heartbeat (left); respiration rate (right)

The duration of the data collection was 60 seconds for each dataset. The digitized data obtained from the ADC is processed in LabVIEW®. The sampled I and Q channels are initially combined using the arctangent function, then the combined data is filtered for respiration rate detection using a low-pass 500 order FIR filter with a cut-off of 0.75Hz and then filtered using a 1000 order band pass FIR filter for heart rate detection using pass-band of 1Hz to 1.5Hz. The Finite Impulse Response (FIR) filters have been designed using a Kaiser Window and Beta value of 6.5 for good main and side lobes characteristics. The extraction of vital signs is done by three different methods: Fast-Fourier-Transform (FFT) in the frequency domain, time-domain autocorrelation with peak detection, and FFT of the autocorrelation output. The filters and DSP to extract vital signs was done in NI’s LabVIEW. The GUI displays the sampled data and results and logs all data (Ichapurapu et al., 2009a).

3.3.3 Third generation noncontact vital signs sensor system

In order to improve the portability of the system, we designed our third generation system to miniaturize the system further (Ichapurapu et al., 2009b). Compared to our earlier system, we made the following changes in this 3rd-gen. design: make the entire system into a single PCB as shown below; the duration of data collection is reduced to 30 sec (from 1 minute); implemented a pass band filter from 0.75 to 1.75 Hz (instead of 1-1.5 Hz) for the calculation of the heart rate. In both cases, the demodulated data was then analyzed and vital signs determined using all three methods mentioned above. For the third generation system the setup consisted of a single PCB based system as shown in Fig. 16 using YAGI antennas. The

Doppler measured vital signs data is shown in Figs. 17-18, which indicate very good data quality. A part of the GUI interface is shown in Fig. 19.

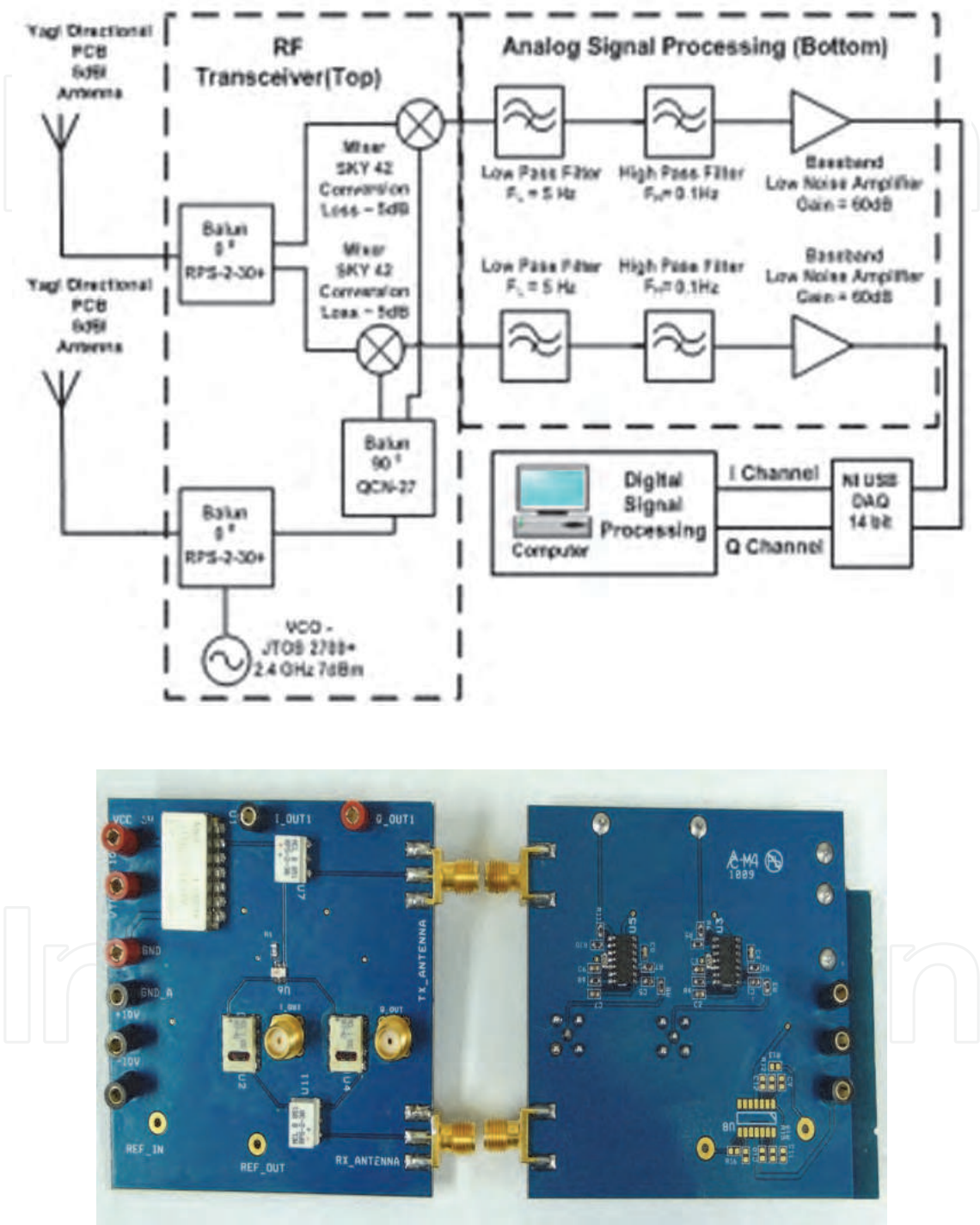


Fig. 16. Third Generation noncontact vital signs sensor system: system block diagram (TOP); single PCB sensor system (Bottom Left: RF PCB; Bottom Right: Analog PCB)

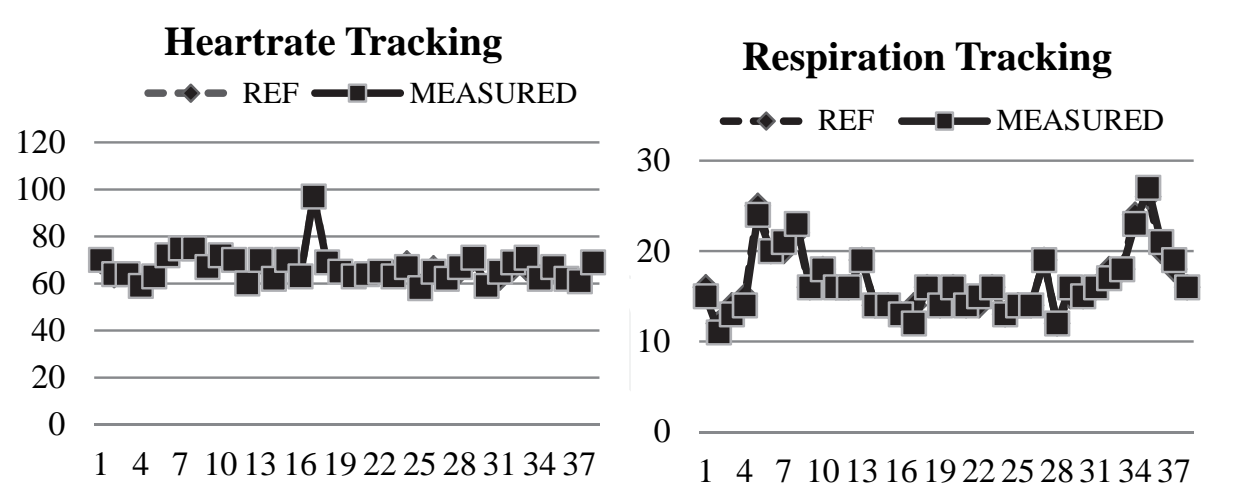


Fig. 17. Data showing how well our 3rd Generation non-contact vital signs sensor system tracks the reference for Heartbeat rate (left); Respiration rate (right)

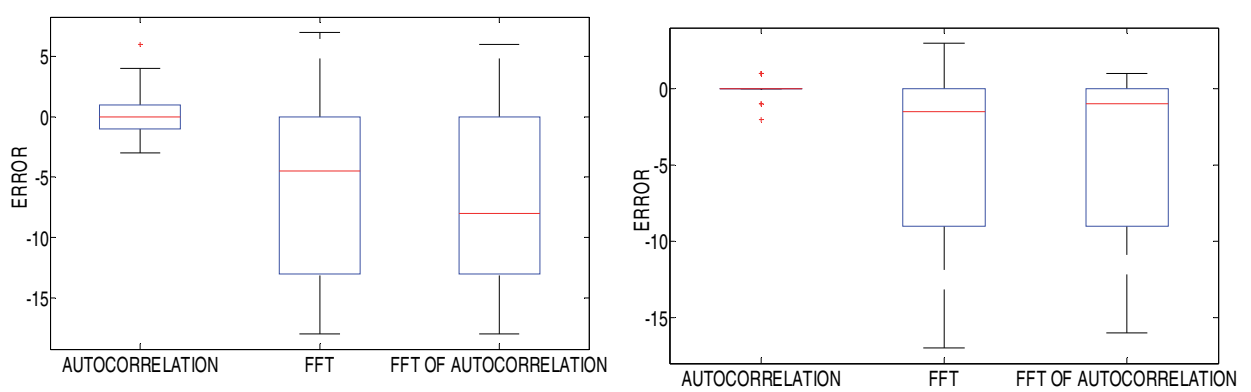


Fig. 18. Box plots showing the measured error distributions from our 3rd Generation non-contact vital signs sensor for the heartbeat rate (left) and respiration rate (right)

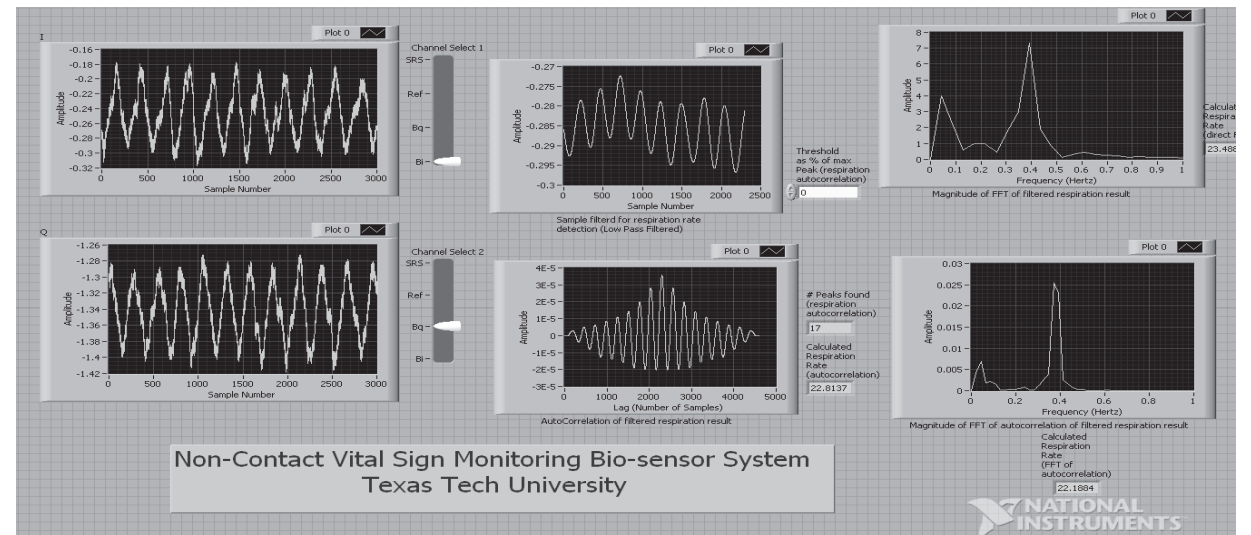


Fig. 19. A screenshot of our developed LabVIEW® GUI for the non-contact sensor system

4. Conclusions and future work

The Doppler radar non-contact system was used to measure the vital signs of a seated subject in a controlled environment. Data was collected using 4 different configurations including a single channel system with discrete RF components, and three different quadrature configurations. Good data accuracy was observed in the extracted vital signs. Quadrature transceiver with arctangent combination and autocorrelation showed the best accuracy for both respiration and heartbeat rate determination. Evaluating performance of the system using both the mean and standard deviation of these results gives us a good indication of the robustness of the system (using only the mean error can be misleading as the error in measurement can be positive and negative). As expected, the mean and standard deviation of the error increase as the distance of the subject from the sensor increases from 0.5m to 1.5m, due to the smaller SNR. Autocorrelation is a solid method with lowest extracted average error < 1 beat/min for respiration rate and average error < 3 beat/min for heart rate for 3rd generation sensor for most cases. Errors in extracted respiration rate are significantly smaller than those of the heart rates due to its larger signal strength. The continuous vital signs data measured from these portable sensors can be wirelessly transferred to healthcare professionals to make life saving decisions and diagnosis of symptoms. For future work, further miniaturization and performance improvement are needed on DC offset cancellation, and resolving the effects of the harmonics of respiration signal overpowering the heartbeat signal. We are working on making the algorithm more robust with some adaptive filtering techniques, where one might estimate the cutoffs for the heartbeat rate detection based on the respiration rate. Our vision is that a continuous log of these vital signs info can also be used to remotely monitor and gauge the recovery of patients, and even for the prevention or prediction of severe illnesses and complications. We also have to go through a couple of clinical trials to evaluate our non-contact vital signs sensor system in terms of false-positives and robustness in a clinical environment. Features such as RFID, motion-detection, etc. can also be integrated in the non-contact vital signs sensors.

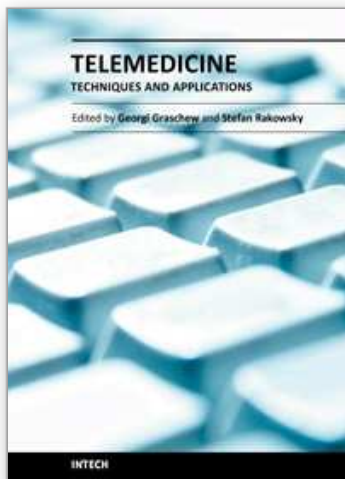
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