Integrated Radar and Communications based on Chirped Spread-Spectrum Techniques

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Abstract—Linear Frequency Modulated (LFM) signals and the associated pulse compression techniques are attractive in applications where highly secure and robust communication is needed. This paper investigates the novel integration of radar and communications utilizing LFM waveforms. The simulations suggest that the performance of the communications-receiver deviates at most 2 dB from the theoretical probability of bit error for $\pi/4$-differential phase shift keying. The simulated radar receiver-operating characteristics for false-alarm probabilities between 10^{-2} and 10^{-4} also compares very well with the theoretical limits for a coherent system.

Index Terms—Chirp signals, radar, spread spectrum communication, wireless communications, co-site interference.

I. Introduction

Multifunctionality in military RF subsystems will reduce cost, minimize the radar cross section and probability of intercept, and allow systems to work simultaneously with tolerable co-site interference in the time domain. Recent attempts at military multifunctional systems have largely been targeted at the concept of broadband RF apertures that are capable of simultaneous operation of communications, radar, and electronic warfare [1]-[2]. The present research takes a different approach, utilizing independent waveform generation and timing control circuits for each function, while allowing the elements of the antenna array to simultaneously transmit and receive communications and radar data. Opposite-slope linear frequency modulated (LFM) waveforms are used for radar and communications pulses. LFM or "chirp" signals have historically been associated with radar, having helped engineers significantly improve the trade-off between range resolution and maximum search range. Interference rejection and robustness in multipath fading environments, which are inherent properties of spread spectrum systems [3], also make chirp signaling very attractive for the expanding wireless communications market. Winkler proposed the use of chirp signals for analog communication in 1962 [4]. Applications to digital communications soon followed [5]-[9].

LFM waveforms have recently been used to develop an indoor spread spectrum communication system [10]-[12]. The system was implemented with two different modulation techniques, first, binary orthogonal keying (BOK), and second, $\pi/4$-differential phase shift keying. BOK exploits the quasi-orthogonality of an 'up-chirp' (frequency increasing with time) and 'down-chirp' (frequency decreasing with time) signal to convey the binary data stream with a '1' corresponding to an up-chirp and a '0' corresponding to a down-chirp. However, BOK modulation is inefficient, because one chirp pulse represents a single bit. In order to increase the data rate the chirp pulses must overlap in time, but this leads to inter-bit interference as described in [10]. To achieve a practical data rate and avoid phase synchronization problems in the communications receiver, the present work utilizes the $\pi/4$-DQPSK method to encode the digital communications data stream.

II. System Architecture

A. Signal Generation and Timing

Surface Acoustic Wave (SAW) devices are implemented for the important signal processing functions of chirp generation and compression. SAW devices are commonly used to generate and compress chirp signals because they are passive, inexpensive and easy to fabricate [6],[13]. However, SAW devices suffer from a rather large insertion loss on the order of 20-30 dB. The integrated architecture employs two SAW chirp filters having identical center frequencies but opposite polarity chirp rates, -40 MHz/μs for the radar and +40 MHz/μs for the communications. In typical phase shift keying communication systems, the Intermediate Frequency (IF) signal is phase modulated by the incoming symbol stream. In the present system, the RF pulse stimulating the chirp filter is phase modulated. The communication bit rate is 1 MHz, which translates to a symbol rate of 0.5 MHz and chirp duration of 2 μs. The radar Pulse Repetition Interval (PRI) is 6 μs and the transmitted pulse width is 2 μs. In each PRI, one up-chirp communications pulse overlaps in time with one down-chirp radar pulse, while the two-subsuquent communication chirp pulses are unperturbed. Fig. 1 shows the radar chirp pulse, the communications chirp pulses, and the overlap between the radar and communications pulses for a 6 μs pulse repetition interval (the real part of the chirp signal envelopes are shown; therefore an up-chirp and down-chirp look similar).

III. Integration Issues

Fluctuations in the amplitude of the signal envelope caused by the overlap of the radar and communications signal at the onset of each PRI (see Fig. 1c) can cause intermodulation products in nonlinear transmitter devices. Distortions caused by the power amplifier have been studied previously in [12].

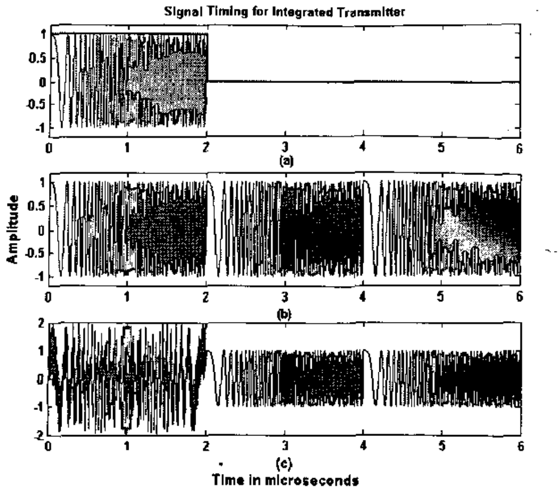

Fig. 1. (a) The real component of the radar chirp complex envelope. (b) The real component of the communications chirp complex envelope. (c) The overlap of the radar and communications chirp pulses

Rapp's model was applied to the data and the analysis clearly shows that at low data rates (< 10 Mb/s), the output backoff levels required for power amplifier linearization are quite low. Therefore, "chirp-based communications show good resistance against nonlinear distortions." This paper also evaluates the effects of the power amplifier (Triquint TGA9083-SCC) on the bit error rate of the integrated system, as discussed in Sec. V. A key issue in the design of the integrated architecture is the isolation between transmitter (Tx) output and the receiver (Rx) input. Isolation on the order of 50-60 dB between transmit and receive paths is typically provided by duplexers in communications receivers by separating Tx and Rx frequencies by 50 MHz or more. However, Doppler shifts commonly observed in radar are typically less than ~10 KHz at microwave frequencies. Therefore, the typical separation of frequencies in duplexer operation is not practical for radar function. To overcome the isolation requirements we have implemented separate receivers for radar and communications as shown in Fig. 2. Self-jamming and mutual interference due to signal waveform design is discussed further in Sec. IV.

TABLE I TRANSMITTER COMPONENT SPECIFICATIONS

Parameter	Chirp Filter	Mixer	Power Amp	Antenna
Center Freq	340 MHz	10 GHz	10 GHz	10 GHz
Bandwidth	80 MHz		5 GHz	
Chirp Rate	40MHz/μs			
Gain	0 dB	-5 dB	19 dB	5 dB
P1dB			39 dBm	
PSAT			40 dBm	
Noise Figure		5dB	10 dB	

IV. SIGNAL TRANSMISSION AND RECEPTION

After chirp signal generation, the combined signal is up-converted using a commercially available mixer (MA/COM MY77) to a frequency of 10 GHz, power amplified, and transmitted thru the additive white Gaussian noise channel via an ideal antenna with a gain of 5dB. In the communications receiver, the signal is received by an ideal antenna with a gain of 5dB, amplified, down-converted and then match-filtered. The match-filter output is then directed to an incoherent π/4-differential demodulator. The radar receiver also obtains the signal via an ideal antenna with 5dB of gain, which directs the signal to a low noise amplifier. The signal is then directed to a down-converter, and then to the input of a matched filter. The output of the matched filter is fed to an envelope detector and then to a simple threshold detector for decision-making. Table 1 lists the specifications of the integrated transmitter.

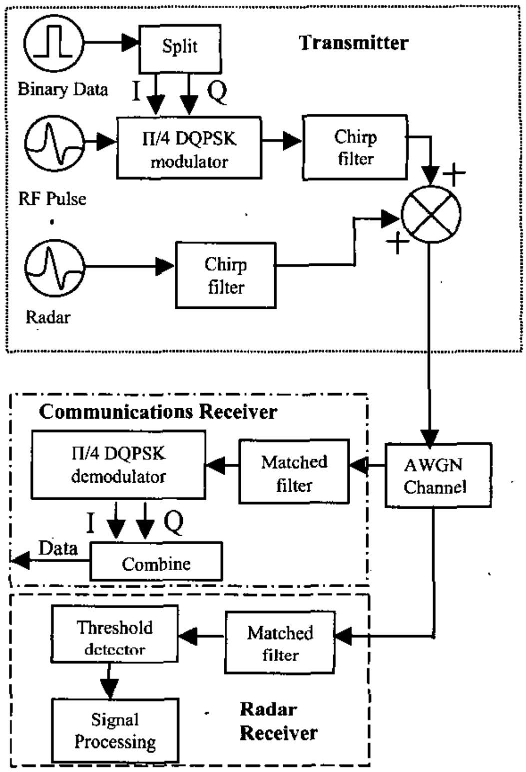

Fig. 2. Schematic block diagram of chirp based integrated radar and communications transmitter and communications receiver.

V. CHIRP SIGNALS

LFM chirp signals are given by the expression

$$x(t) = a(t)\cos(2\pi fot + \mu t^2 \pi + \theta) \tag{1}$$

where $a(t)$ is the signal envelope defined by $a(t)=0$ for $|t| > T/2$, with T being the duration of the chirp signal, fo the center frequency, μ the chirp rate, and θ a phase constant. As mentioned earlier an up-chirp will have $\mu > 0$ and a down-chirp will have $\mu < 0$. The chirp bandwidth, B, which is centered at fo, is defined as the total range of the instantaneous frequency

$$B = |\mu| T \tag{2}$$

The output of a filter matched to a linear chirp signal, is given by the autocorrelation function [5]

$$g(t) = \sqrt{BT}\,\frac{\sin \mu \frac{t}{2}(T - |t|)}{\mu t}\cos(2\pi f o t) \qquad (3)$$

for $-T < t < T$. This assumes that $a(t)$ is a rectangular function (applicable to large time-bandwidth products) centered at t = 0 with an amplitude of 1, and a duration of T. The envelope of the autocorrelation function has its maximum value $(BT)^{1/2}$ at t = 0, and its first zeros at $t = \pm 1/B$. The transmitted chirp signal is thus compressed in time by its matched filter. A figure of merit often used for spread spectrum systems is the compression or processing gain, given by

$$P_g = TB \qquad (4)$$

The processing gain is a good measure of the resistance to jamming, noise, and other interference effects in the receiver.

VI. Signal Orthogonality

To avoid mutual interference the processing gain should be as large as possible and the radar and communications signals should be orthogonal. Theoretically, any two-chirp signals that span non-overlapping frequency regions are mutually orthogonal, which for chirp signals i and j is stated mathematically as [14]

$$S_i(f) \bullet S_j^*(f) = \begin{cases} c, i = j \\ 0, i \neq j \end{cases} \qquad (5)$$

where $S_i(f)$ is the Fourier transform and $S_j^*(f)$ is the conjugate of the Fourier transform and $\bullet$ denotes multiplication. Practical chirp signals with finite time duration will not satisfy Eq. (5) exactly. However, one can approximately meet the orthogonal condition by maximizing the ratio of the side lobe levels of the autocorrelation output to the output of the cross-correlation function. This ratio can be enhanced by using opposite polarity for the frequency sweep of the two signals, and by completely separating the two signals in frequency space [14]. In our system, the radar and communications signals occupy the same frequency band, but have opposite polarity in their respective frequency sweeps. Interference rejection is often measured by the processing gain as mentioned above. Our system has a relatively large processing gain of 160, or 22 dB when compared to the IEEE 802.11 Standard of at least 10 dB.

VII. System Simulations

A. *Communication Simulation Results*

Co-simulations have been carried out using the MATLAB and Serenade software programs to evaluate the performance of the integrated architecture with a sample size of 10^5 bits. Two case studies were analyzed. The first simulation utilized ideal components in the transmitter and receiver. The second used S-parameters to describe the non-ideal mixer and power amplifier in the transmitter. In each simulation, the chirp filters and the LFM signals generated were assumed to be ideal. The channel was modeled as an additive white Gaussian noise channel with a free space link characteristic given by:

$$S_{21} = S_{12} = \frac{\lambda}{4\pi R}; \quad S_{22} = S_{11} = 0 \qquad (6)$$

The communications receiver was ideally modeled in both simulations so that the carrier-to-noise ratio at the antenna will be equal to the signal-to-noise ratio at the detector. In Fig. 3, the communication receiver simulation results are plotted as bit error rate versus signal-to-noise ratio. As the figure shows, the BER of the communications receiver compares very favorably with the theoretical BER for π/4-DQPSK modulation. The ideal transmitter and receiver BER was 0 at SNR greater than 11dB. Practical modeling of the power amplifier and mixer by their respective S-parameters introduced a few bit errors, which vanished for signal to noise ratios greater than 12 dB. The power delivered to the antenna is approximately 10 W. The theoretical probability of bit error is given by [15]

$$Pe = Q(a,b) - \frac{1}{2}I(a,b)\exp(\frac{a^2+b^2}{2}). \qquad (7)$$

In this expression $Q(a,b)$ is the Marcum's Q function, $I(a,b)$ is the modified Bessel function of the first kind and zero order, and a, b are defined

$$a, b = \sqrt{SNR}\left(\sqrt{2+\sqrt{2}} \mp \sqrt{2-\sqrt{2}}\right) \qquad (8)$$

Integrating over the symbol duration and sampling according to a clock recovery scheme may improve bit detection even for an indoor environment where harsh ISI is observed [11].

B. *Radar Simulation Results*

Radar simulations were conducted using S-parameters to describe the transmitter power amplifier, mixer, and free-space link. The chirp signals, filters, and the radar receiver were all assumed to be ideal. However, the received signals are dependent on the target range and radar cross-section (RCS). The channel is modeled as an additive white Gaussian noise channel, and the free-space link transmission is given by:

$$_{21} \qquad _{12} \qquad \sqrt{\frac{}{(4\ )^3 R^4}}$$

$$S_{22} = S_{11} = 0 \qquad (9)$$

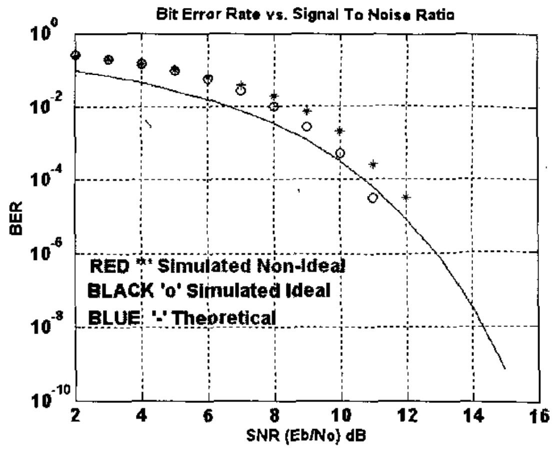

Fig. 3. Simulated BER versus SNR for the π/4-DQPSK communications receiver.

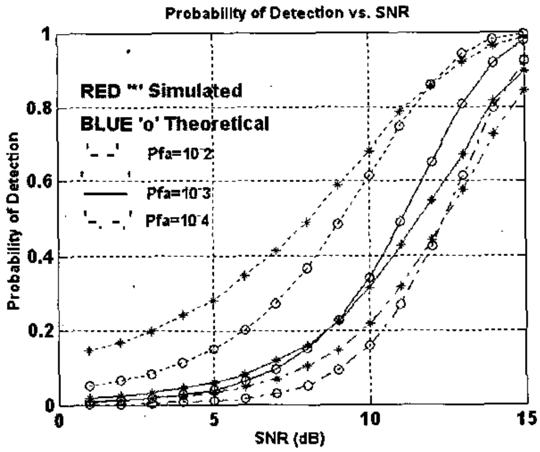

Fig. 4. Radar ROC diagrams.

where σ is the radar cross section, λ is the operating wavelength and R is the target range). The threshold for detection was determined by [16]:

$$V_t = \sqrt{2\psi^2 \ln(1/Pfa)} \tag{10}$$

where ψ is the root mean square noise voltage. The ROC diagram shown in Fig. 4 is for a target with an RCS of 1 m^2. The probability of detection is given by:

$$P_d = Q[\sqrt{SNR}, \sqrt{2\ln(1/pfa)}] \tag{11}$$

Where Q is the Marcum's Q function. The figure shows that the P_d vs. Pfa curves are close to theoretical limit established by the physical noise ψ.

VIII. Conclusions

We have presented and analyzed a multi-functional RF architecture that integrates a simple search radar and digital communications function using LFM signaling. Orthogonality between the waveforms was enhanced by using separate up- and down chirps for the radar and communications pulses, respectively. Π/4-DQPSK modulation was also used to achieve a high data rate and to eliminate the difficulties of phase synchronization in the receiver.

Simulations in an additive white Gaussian noise channel reveal that the BER performance relative to theoretical π/4-DQPSK systems is very good. And the receiver operating characteristics in the radar function are satisfactory. Therefore, the use of LFM waveforms to convey radar and digital voice or data from a common transceiver is a promising multifunctional approach that uses signal processing rather than hardware to mitigate the mutual interference problem.

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