MULTICHANNEL RADAR BACKSCATTER COMMUNICATION AND LOCALIZATION

Itay Cnaan-On^{*} Stewart J. Thomas^{*} Matthew S. Reynolds^{*†} Jeffrey L. Krolik^{*}

* Department of Electrical and Computer Engineering, Duke University, Durham, NC, USA [†]Department of Electrical Engineering, University of Washington, Seattle, WA, USA

ABSTRACT

This paper concerns the use of wideband RF backscatter from semi-passive RF tags for energy-efficient wireless telemetry. Using LFM (Linear Frequency Modulated) signals from a radar basestation, we present a method for joint ranging and communications with a distributed set of sensor nodes. Backscatter signaling from each node results in modulation on a sub-carrier frequency determined by the node distance from the radar. Upper bounds on communications rate per coverage area are presented, as well as clutter filtering to suppress ground components. Experimental results from a brass-board microwave system in an indoor environment are presented and discussed.

Index Terms— radar communication, localization, radio frequency identification, microwave communication

1. INTRODUCTION

Backscatter modulation has been shown to greatly increase the wireless communication power efficiency of remote devices and sensors. This modulation technique allows a device to wirelessly telemeter information without operating a local transceiver. Instead, devices implementing backscatter modulation rely on the incident EM signal for the RF carrier and convey information within controlled reflections [1, 2]. This allows for simplified circuitry and a reduction in required power of the remote device at the expense of increased complexity, processing and power at the base station. Communication using backscatter modulation is widely used in the field of Ultra High Frequency (UHF) Radio Frequency Identification (RFID), and is especially well-suited for sensor communication where ultra-low power and extremely simple design are key operating parameters. Recent work has demonstrated the potential for backscatter-based sensors intended for bio-signal recording [3, 4, 5], logistics/asset monitoring [6] and environmental sensing [7].

Typically, backscatter systems use (single-frequency) continuous wave (CW) transmission waveforms for data interrogation. While sensor data can be contained within

amplitude and/or phase changes of the backscattered signal [8, 9], it is difficult to distinguish range dependent phase and amplitude changes (used for extracting localization information) from clutter fluctuations within the environment [10]. It is likewise difficult for a single reader to separate simultaneous data streams when multiple tags are talking. Existing protocols for UHF RFID systems require only one tag to talk while others are silent. Previous work have used antenna array processing [11], or synchronization protocols [12] to overcome these limitations. Other work has examined the localization capabilities of using linear Frequency Modulated Continuous Wave (FMCW) with RF tags [6], or communication for a single active RF tag [13].

In this work, we explore the use of a LFM FMCW radar waveforms for ranging and multichannel data transfer from multiple semi-passive backscatter modulators intended for long-term, wide-area sensing. The processing method we describe allows for streaming data from multiple sources provided transmission rates fall within an upper bound that is derived. In addition to the joint signal processing framework for localization and communication, we employ coding-based clutter filtering to aid in discriminating the signal of interest from clutter return. Results from an indoor experimental analysis verifying the processing are provided.

2. SIGNAL MODEL

The radar transmits linear FMCW 'chirp' pulse train x(t). The sweep rate β of each chirp pulse determines the level of instantaneous bandwidth of the signal and is measured in units of Hz/s. A single chirp pulse has the structure of

$$x_p(t) = e^{j2\pi f_c t + j\pi\beta t^2} \tag{1}$$

with f_c representing the center carrier frequency, $\beta = B/T_p$ denoting the sweep rate, B denoting the chirp pulse bandwidth and T_p being the chirp pulse duration. The chirp pulse train is

$$x(t) = \sum_{n=-\infty}^{\infty} \operatorname{rect}\left[\frac{t-n \cdot T_p}{T_p}\right] \cdot x_p \left(t-nT_p\right)$$
(2)

where rect[t] is the rectangular function. In the remainder of this document, the term for time limiting each chirp pulse using $rect[\cdot]$ is omitted for readability.

This work is supported by the Office of Naval Research (ONR) under grant number N00014-13-1-0065. This work is also supported by the National Science Foundation (NSF) Wireless Intelligent Sensor Networks (WISeNet) program under grant number DGE-1068871



Fig. 1. Joint processing framework for localization and communication

The node structure is extremely simple and consists of an antenna that is switched between two loads, open/short circuit, corresponding to two constellation symbol states as can be seen in Fig. 2. A portion of the incident wave absorbed by the antenna is reflected at the load and re-radiated back to the signal source. By changing the load connected to the antenna, the reflected fields also change allowing for control of the data symbols using a single switch. Modulating states between open/short circuit creates binary phase shift keying or pulse amplitude modulation (BPSK/PAM) over a rectangular pulse train. The message signal driving the node switch is expressed as

$$m(t) = \sum_{k=-\infty}^{\infty} a_k \cdot p\left(t - kT_b\right) \tag{3}$$

where $a_k \in \{+1, -1\}$ is the binary phase shift for symbol k and p(t) is a rectangle pulse symbol with duration T_b representing a single bit period.

The backscatter reflecting from node i is received at the radar with time delay $\tau_i = 2r_i/c$ where c is the speed of light and r_i is the one-way range between node i and the radar. For the scope of this document, it is assumed that the node is stationary (has no Doppler frequency shift). The resulting signal from node i received at the radar is

$$s(t, r_i) = \alpha(r_i) \sum_{n} \sum_{k} \{ \overbrace{x_p(t - nT_p - \tau_i)}^{\text{chirp pulses}} \cdot \overbrace{a_k p(t - kT_b - \frac{\tau_i}{2})}^{\text{node symbols}} \}$$
(4)

where $\alpha(r_i)$ is the attenuation resulting from the range $2r_i$ the signal travels.

The transmitted chirp signal is also reflected from clutter or stationary elements in the surroundings (such as ground, walls etc.). The clutter can be modeled as a collection of point scatterers at different ranges and the cumulative clutter return at the radar is

$$c(t) = \sum_{c \in C} \sum_{n} \alpha(r_c) x_p (t - nT_p - \tau_c)$$
(5)



Fig. 2. Simplified operation of radar base station and backscatter node

with C representing the set of clutter points and $\tau_c = 2r_c/c$ is the time delay associated with the range of a clutter point from the radar r_c .

Other nodes are also operating simultaneously at different ranges

$$\sum_{g \in G} s(t, r_g) \tag{6}$$

where G is the set of all operating nodes and $s(t, r_g)$ is the backscatter from a node residing at range r_q .

The cumulative signal at the radar is therefore a sum of all modeled signals with additive white noise, resulting in

$$r(t) = s(t, r_i) + c(t) + \sum_{g \in G} s(t, r_g) + n(t)$$

$$= \sum_{i \in G} \alpha(r_i) \sum_n \sum_k \left\{ x_p \left(t - nT_p - \tau_i \right) \cdot a_{k,i} p \left(t - kT_b - \frac{\tau_i}{2} \right) \right\}$$

$$+ \sum_{c \in C} \alpha(r_c) \sum_n x_p \left(t - nT_p - \tau_c \right) + n(t)$$
(7)

where $a_{k,i}$ is the phase shift for symbol k of node i.

3. LOCALIZATION PROCESSING

The high resolution range processing of the node is achieved by pulse compression (sometimes referred to as 'de-chirp' or 'de-ramp'). This process computes the crosscorrelation between the received signal r(t) and the transmitted chirp pulse signal x(t) in the time domain [14]

$$r_d(t) = \langle r, x \rangle(t) = \int_{t'=-\infty}^{\infty} r(t') x^*(t+t') dt'$$
(8)

The outcome of the de-chirp process (derivation omitted for space constraints) is

$$r_{d}(t) = \sum_{i \in G} \alpha(r_{i}) \sum_{n} \sum_{k}$$
(9)
$$\underbrace{\left\{ \overbrace{a_{k,i}p(t-kT_{b}-\frac{\tau_{i}}{2}-\delta_{i})}_{\text{node symbols}} \cdot \overbrace{e^{j2\pi\beta\tau_{i}(t-nT_{p})}}^{\text{range induced subcarrier}} \right\}}_{\substack{clutter point induced tone}}$$

$$+ \sum_{c \in C} \alpha(r_{c}) \sum_{n} \overbrace{e^{j2\pi\beta\tau_{c}(t-nT_{p})}}^{\text{clutter point induced tone}} + n'(t)$$



Fig. 3. Measured range-Doppler processing output of experimental data at the reader side. The highlighted area shows a node signaling at a low-rate at range of 4m

where τ_i is node *i* delay, δ_i is clock offset between the radar clock and the node *i* clock and n'(t) is noise modeled as an additive white Gaussian noise (AWGN) process.

For a return at some time delay τ_i , the de-chirp processing associates a corresponding induced tone whose frequency f_i is a function of that time delay $f(\tau_i)$. The exact range of the node of interest can then be easily recovered by using the relation

$$f_i = \beta \cdot \tau_i = \frac{B}{T_p} \frac{2r_i}{c}.$$
 (10)

In the case of a singular scatterer point (such as a clutter point or a non backscattering signaling node), a single tone will be present corresponding to that time delay. However, in the case of a backscattering signaling node, the expected outcome would be a time-domain multiplication (product mixer) of the the node signal m(t) and the range induced subcarrier $f_i = e^{j2\pi\beta\tau_i t}$ as is demonstrated in (9). In terms of range processing, instead of a single frequency corresponding to a single range point, the backscattering signaling node will appear in the de-chirp outcome as a bandpass process centered around f_i . For the purposes of maximizing range estimation accuracy, a low-rate backscatter preamble signal (e.g. 101010), relative to the waveform repetition frequency, is adequate for separating the node return from the ground clutter while maintaining high SNR within a radar range bin. An example of a range Doppler surface for a backscattered signal used for localization is shown in Fig. 3. Note the sidebands of the node are clearly visible.

4. COMMUNICATION PROCESSING

4.1. De-chirp process

The first step is similar to the step taken in range processing, which is a cross-correlation between the received and transmitted signal. For a return at some time delay τ_i , the dechirp processing will associate a corresponding induced tone whose frequency f_i is a function of that time delay $f(\tau_i)$. For a transmitting node *i*, the expected outcome would be a timedomain multiplication (product mixer) of the the node signal $m_i(t)$ and the range induced carrier $f_i = e^{j2\pi\beta\tau_i t}$ (see (9)).

4.2. Estimation of the range induced carrier

The next step is to isolate and process each transmitting node and to recover its individual spectrum. Given a sufficiently accurate estimate of the node range from the training preamble, the carrier frequency, f_i , can be used to demodulate the communications signal. A conjugated tone $e^{-j2\pi\beta\tau_q t}$ such that $\tau_q = \tau_i$ can then be mixed and the outcome passed through low pass filter (cut off at $|f| \leq \frac{1}{T_b}$) to recover the original node signal in base-band.

$$r_{b}(t) = r_{d}(t) \cdot e^{-j2\pi\beta\tau_{q}(t-nT_{p})} = \alpha(r_{i}) \sum_{n} \sum_{k}$$
(11)
$$a_{k}p(t-kT_{b}-\frac{\tau_{i}}{2}-\delta_{i}) \cdot e^{j2\pi\beta(\tau_{i}-\tau_{q})(t-nT_{p})}$$
$$\underset{\tau_{q}=\tau_{i}}{=} \alpha(r_{i}) \sum_{n} \sum_{k} a_{k}p(t-kT_{b}-\frac{\tau_{i}}{2}-\delta_{i})$$

Note that range estimation error in the demodulation process can be expected to have a similar impact as carrier frequency uncertainty in conventional systems.

4.3. Match filtering for the node symbols

The next step is now recover the symbols from the output term after de-mixing the range induced carrier in (11). The optimal linear filter to maximize the SNR in the presence of white noise is the matched filter for the rectangular pulse symbol

$$r_{MF}(t) = \int r_b(t-u)p^*(-u)du$$

$$= \alpha(r_i)\sum_n \sum_k a_k \int p(t-u-kT_b - \frac{\tau_i}{2} - \delta_i)p^*(-u)du$$

$$= \alpha(r_i)\sum_n \sum_k a_k \operatorname{tri}(t-kT_b - \frac{\tau_i}{2})$$
(12)

where p(t) is a rectangular pulse, we define the triangle function tri as the convolution of two rectangular functions rect and assume the clock offset δ_i between the radar and the node is known.

Methods of clock recovery for asynchronous communication are currently under investigation.

5. UPPER BOUND ON NODE DATA RATE AND NODE SPATIAL DENSITY

The outcome of the de-chirping process developed in section 4 is that transmitting nodes signal is product-mixed with a range induced subcarrier corresponding to the node range. Therefore, every node transmitting at symbol rate $1/T_b$ will be centered in frequency around f_i as can be seen in Fig. 4. The question which arises is how close in range can two neighboring nodes be without causing aliasing to each other



Fig. 4. Measured spectrum of received signal before (red) and after (blue) clutter filtering. Node signal is centered around subcarrier of $f_i = 40Khz$ (Parameters used: $f_c = 2.45Ghz$, B = 40Mhz, $T_p = 20ms$, $T_b = 40ms$)

in the de-chirped frequency domain. More formally we can define nodes spaced apart by range Δr . Their respective frequency spacing will be $\Delta f_r = \beta \Delta \tau = \frac{B}{T_p} \frac{2\Delta r}{c}$. We assume node 1 transmits at symbol rate $\frac{1}{T_{b_1}}$ and that node 2 transmits at symbol rate $\frac{1}{T_{b_2}}$. In order to prevent override or aliasing, the cumulative symbol rate has to satisfy

$$\frac{B}{T_p} \frac{2\Delta r}{c} \ge \left[\frac{1}{T_{b1}} + \frac{1}{T_{b2}}\right]$$
(13)

This bound provides a fundamental trade-off relations between node data rate, range separation and the bandwidth assigned for the LFM carrying waveform. In other words, a node transmitted data rate will be bounded by the amount of spatial difference with other nodes which are simultaneously transmitting. Also, the larger the bandwidth is assigned for the LFM carrying waveform, the higher the data rate the nodes can use without causing aliasing to neighboring nodes.

6. CLUTTER FILTERING USING LINE CODING

6.1. Clutter statistics analysis

In most cases, the amount of energy reflected back from clutter is much higher than the amount of energy reflected from nodes. This makes it harder to discriminate between the signal of interest and the clutter masking it. Clutter can be modeled as a set of scatterer points that will convert after de-chirping respectively into a collection of tones. Assuming clutter points are stationary (have near zero Doppler) and processing time short enough, then the output of the de-chirp process clutter can be viewed as an unknown process (a sum of collection of tones and amplitudes) which is periodic with the chirp pulse rate $1/T_p$, since clutter returns will be the same across chirp pulses. By exploiting this clutter statistics, the clutter can be significantly mitigated.

6.2. Node's line coding to mitigate clutter

A line coding and decoding technique can be used such that the decoding process exploits the known clutter statistics and filter it, yet let the decoded information symbols of the node to be passed through. The idea is that the node sends random set of symbols, which fills up a whole duration of chirp pulse (it can be assumed that the chirp pulse duration



Fig. 5. Measured received signal after removal of range induced carrier and clutter (blue) and expected (green) (Experimental results using same parameters as in Fig. 4)

 T_p is a design parameter known in advanced to the nodes). Then on the following chirp pulse, the node sends a negated version of the same set of symbols. The coding then continues similarly for the next pair of chirp pulses. The reader decodes each two received chirp pulses by subtracting the second from the first. In that way, the clutter is removed almost completely but the decoded symbols are now recovered. For example consider the case of sending a set of 4 information symbols by encoding and decoding (by subtracting the second chirp return from the first chirp return and removal of the clutter):

$$\overbrace{0100}^{\text{first chirp second chirp}} 10111 \Longrightarrow 0100$$

This method is similar in concept to Moving Target Indication (MTI) method [14] used to discriminate a moving target from clutter in the radar domain, but is employed in this work for the purpose of communication.

Experimental results which include the clutter filtering are shown in Fig. 4. The spectrum of the de-chirped RF signal is shown by the red trace. The most dominant components are the spectral peaks corresponding to the clutter from the stationary surroundings. The node is masked by the clutter but can be recovered by differencing waveform repetition intervals as discussed. The resulting communication signal obtained after differencing is shown by the blue trace. Fig. 5 shows the time domain baseband outcome after removing the range induced tone. The processed signal follows closely the expected decoded node signal. Since it is low pass filtered and pulses have infinite frequency capacity, some ripples are noticeable.

7. CONCLUSION

In this work we have provided the theoretical analysis for using FMCW waveforms for communication and localization of semi-passive RF nodes intended for long-term widearea deployment. Fundamental bounds relating the spatial density of sensors and their symbol rate are derived, along with an analysis of the localization accuracy. In addition, we have presented two key processing methods: The first allows multiple nodes to transmit simultaneously without crossinterference, and the second proposes an efficient clutter mitigation/channel equalization technique using line coding. Experimental results that validates the processing methods were presented.

8. REFERENCES

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