Quadrature Amplitude Modulated Backscatter in Passive and Semi-Passive UHF RFID Systems

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Abstract—Passive and semi-passive UHF RFID systems have traditionally been designed using scalar-valued differential radar cross section (DRCS) methods to model the backscattered signal from the tag. This paper argues that scalar-valued DRCS analysis is unnecessarily limiting because of the inherent coherence of the backscatter link and the complex-valued nature of load-dependent antenna-mode scattering from an RFID tag. Considering modulated backscatter in terms of complex-valued scattered fields opens the possibility of quadrature modulation of the backscatter channel.

When compared with binary ASK or PSK based RFID systems which transmit one bit of data per symbol period, tags employing vector backscatter modulation can transmit more than one bit per symbol period. This increases the data rate for a given on-chip symbol clock rate leading to reduced on-chip power consumption and extended read range. Alternatively, tags employing an $M$-ary modulator can achieve $\log_2 M$ higher data throughput at essentially the same DC power consumption as a tag employing binary ASK or PSK.

In contrast to the binary ASK or PSK backscatter modulation employed by passive and semi-passive UHF RFID tags, such as tags compliant with the widely used ISO18000-6c standard, this paper explores a novel CMOS-compatible method for generating $M$-ary QAM backscatter modulation. A new method is presented for designing an inductorless $M$-ary QAM backscatter modulator using only an array of switched resistances and capacitances. Device-level simulation and measurements of a 4-PSK/4-QAM modulator are provided for a semi-passive (battery-assisted) tag operating in the 850-950 MHz band. This first prototype modulator transmits 4-PSK/4-QAM at a symbol rate of 200kHz and a bit rate of 400kbps at a static power dissipation of only 115 nW.

Index Terms—QAM backscatter, Passive RFID tags, CMOS integrated circuits, Backscatter, Quadrature amplitude modulation, UHF RFID, backscatter phase rotation

I. INTRODUCTION

Ultra-high frequency (UHF) radio frequency identification (RFID) systems employing modulated backscatter communication links, such as those based on the widely deployed ISO18000-6c or EPC Global Class 1 Generation 2 specification [1], have traditionally been analyzed using techniques first developed to analyze radar systems. This paper demonstrates that scalar valued differential radar cross section (RCS) techniques are unnecessarily limiting for RFID system analysis because of the inherent coherence of the backscatter link and the complex-valued nature of load dependent scattering from an RFID tag’s antenna.

In most commonly deployed UHF RFID systems, the UHF backscatter link employs a simple binary modulation scheme, such as ASK or PSK backscatter generated by a two-state modulation of the impedance presented to a transponder’s antenna (Fig. 1). Because tag to reader distance is generally unknown a priori, most readers employ homodyne I/Q demodulation and rotate the received constellation until it falls on the real axis before demodulating as if the signal were ASK regardless of whether the tag’s modulator is configured to generate ASK or PSK modulation. In this paper, we exploit multi-state complex-valued load dependent scattering to yield quadrature amplitude modulation (QAM) backscatter that is compatible with the homodyne reader architecture.

This paper, an expanded version of [2], demonstrates the use of load dependent scattering to generate QAM backscatter with a simple modulator suitable for single-chip CMOS implementation. The modulator presented in this work has a static power consumption of only 115 nW. Related work has presented a vector modulator based on a PIN diode phase shifter controlled by bias currents generated by two digital to analog converters (DACs) [3]. The latter circuit has a static power consumption of 80 mW, which is prohibitively high power consumption for either passive or semi-passive tags. In contrast to the present approach, it is not suitable for monolithic implementation because PIN diodes and wideband quadrature hybrids are not available in standard CMOS processes.

The use of QAM in passive and semi-passive UHF RFID opens up many new avenues for UHF RFID, including higher spectral efficiency than is possible with ASK or PSK and improved data throughput with reduced on-chip symbol clock rate.

II. SCALAR-VALUED RCS IN RFID SYSTEM ANALYSIS

For a monostatic radar illuminating an uncooperative target, such as an aircraft or a ship, the reflected energy from the target is often estimated using the target’s scalar RCS. Scalar
RCS in this context is simply a measure of the target’s structural scattering at a particular attitude with respect to the radar. RCS is defined in terms of incident and scattered fields

\[
\sigma = \lim_{r \to \infty} \frac{4\pi r^2 |E_s|^2}{|E_t|^2}
\]

where \( r \) is the radar to target separation and \( E_s, E_t \) are the scattered and incident electric fields at the target. Assuming free space propagation between the radar and the target, the radar equation is then employed to estimate the magnitude of the received signal due to radar illumination of the target

\[
P_r = \frac{P_T G_T G_R \lambda^2 \sigma}{(4\pi)^3 r^4}
\]

where \( \sigma \) is target RCS as previously defined, \( P_R, P_T \) are signal power at the radar’s receiver and transmitter, \( G_R, G_T \) are the gains of the radar’s receive and transmit antennas, and \( \lambda \) is the radar’s wavelength.

Unlike a typically uncooperative radar target, the goal of a passive or semi-passive UHF RFID transponder or ‘tag’ is to communicate information to the reader by modulated backscatter through a mutually agreed modulation and data format. The backscattered field consists of structural scattering, which is not intentionally modulated, as well as load-dependent scattering that is modulated with data by varying the impedance presented by the tag circuit to the tag’s antenna as shown in Fig. 1. It has been shown that careful load selection can be used to maximize the backscattered field amplitude as discussed in early backscatter work [4] as well as in modern UHF RFID systems [5]–[7]. A typical CMOS implementation of a binary PSK modulator, showing the switched modulating capacitance, appears in Fig. 2.

In an ASK or PSK modulation scheme, the tag’s backscattered power is usually analyzed [8] in terms of its differential RCS, defined by

\[
\Delta \sigma = \frac{\pi \eta^2}{4\lambda^2 R_a^2} |\Gamma^*_1 - \Gamma^*_2|^2 \kappa^2
\]

where \( \eta \) is the medium impedance (usually free-space), \( R_a \) is the real portion of the antenna impedance, \( \kappa \) is a constant containing the antenna equivalent height, and \( \Gamma^* \) is the conjugate match reflection coefficient

\[
\Gamma^* = \frac{Z_a^* - Z_L}{Z_a^* + Z_L}
\]

for resonant antenna impedance \( Z_a \) and complex load impedance \( Z_L \).

By relating the antenna equivalent height to its gain \( G \), we find a more convenient form for scalar-valued differential RCS

\[
\kappa = \frac{G a^2 R_a}{\pi \eta}
\]

\[
\Delta \sigma = \frac{\lambda^2}{4\pi} G^2 |\Gamma^*_1 - \Gamma^*_2|^2
\]

which is the form of differential RCS widely used in the RFID system design community. The differential RCS is frequently substituted into the radar equation (2) to estimate backscatter signal power versus distance.

III. COMPLEX-VALUED SCATTERED FIELDS

Hansen [10] previously analyzed a special case of radar scattering where the scatterer is an antenna connected to a load. This work relates Hansen’s scattering results to the UHF RFID context and extends it to multiple load impedance states.

The reflected field from a tag can be decomposed into structural scattering and load dependent scattering terms

\[
E_{scat}(Z_L) = E_{scat}(Z_a^*) + \Gamma^* E_{ant} I_m I_{ant}
\]

where \( \Gamma^* \) is the conjugate match reflection coefficient previously defined, \( I_m \) is the antenna current when the load is conjugate matched and \( E_{ant} \) is the radiated field when the antenna is driven by current \( I_{ant} \).

By substituting two different complex-valued reflection coefficients from two complex-valued load impedances into (7), we find the complex-valued differential backscatter \( E \) field that is observed for ASK or PSK backscatter

\[
\Delta E_{scat} = (\Gamma^*_1 - \Gamma^*_2) E_{ant} I_m I_{ant}
\]

In contrast to the traditional differential RCS of (6) that yields only scalar results, the differential \( E \) field scattering is a vector that, when demodulated, may occupy any quadrant of the I/Q plane. The critical observation is that applying carefully selected loads to the antenna will result in the scattered \( E \) field exhibiting amplitude modulation, phase modulation, or both simultaneously and independently yielding a QAM constellation.

In the simplest implementation, an \( M \)-ary QAM modulation scheme can be implemented with \( M \) distinct load impedances that are switched across the terminals of the tag’s antenna, although other implementations are possible, including a simple impedance (DAC) where weighted arrays of inductors, capacitors, and resistors are used to reduce the number of lumped elements to fewer than \( M \). In a monolithic implementation this translates to a reduction in die area and thus a cost and yield improvement. For many types of antennas it is possible to design such an impedance DAC with only resistors and capacitors, which makes the approach particularly suitable for CMOS integration.

IV. DEMODULATION IN PASSIVE AND SEMI-PASSIVE UHF RFID SYSTEMS

Current passive and semi-passive RFID tag modulators are designed either to modulate the real part of the tag IC’s reflection coefficient, yielding ASK backscatter, or the imaginary part, yielding PSK backscatter. The most common
circuit implementation of ASK for passive devices switches between a matched state, which maximizes power delivered to the passive tag’s power rectifier circuitry, and a load resistance that introduces a deliberate mismatch to produce a backscatter signal. A binary PSK modulation circuit is shown in Fig. 2 where the modulating transistor switches a capacitance across the antenna’s terminals to introduce a phase shift in the scattered field.

Because typical RFID readers employ homodyne receivers, the backscatter link is coherent to the reader’s transmit local oscillator. Due to this coherence, the phase of the backscatter as observed in the demodulated baseband changes with tag to reader distance. As the tag moves radially outward from a reader, the phase of the backscatter field incident on the receiver rotates at a rate of $2\pi$ radians per half-wavelength of distance. At the 860-950 MHz frequencies typically used for UHF RFID, this leads to a rotation of $2\pi$ every $\approx 16$ cm. Since the reader to tag distance $r$ is usually initially unknown and both ASK and PSK modulation schemes are permitted by most RFID standards [1], the reader must be able to demodulate tag signals arriving at any phase. For binary modulation schemes such as ASK or PSK, the reader’s baseband signal processing software typically rotates the received signal vector from its arrival angle in the I/Q plane back to the in-phase axis prior to data slicing to demodulate the tag’s binary data. Most existing RFID reader hardware is therefore not restricted to binary ASK/PSK backscatter modulation. While upgraded reader baseband signal processing software would be required to demodulate $M$-ary QAM data with some increase in computational complexity, QAM demodulation is supported by the existing RFID reader architecture.

V. RECEIVED SIGNALS IN THE PRESENCE OF MULTIPATH AND TAG MOTION

The $E$ field at the reader’s receiving antenna is composed of three components

$$E_{\text{rcv}} = E_{\text{env}} + E_{\text{scat}}(Z_L) + N$$

where distance-induced attenuation and phase shift of the scattered field from the tag is $E_{\text{scat}}$, $E_{\text{env}}$ represents scattering of the transmitted signal from stationary objects in the environment, and $N$ represents all environmental sources of noise, including scattering of the transmitted signal due to reflective objects in the environment. After expanding this expression into its constituent parts, we note that the received signal $E$ field

$$E_{\text{rcv}} = E_{\text{env}} + E_{\text{scat}}(Z_L^*) + \Gamma^* \frac{E_{\text{env}I_m}}{I_{\text{ant}}} + N$$

(10)

can be treated as a sum of vector components

$$E_{\text{rcv}} = \vec{A} + \vec{B} + \vec{C} + \vec{N}$$

(11)

decomposed of contributions from environmental multipath ($\vec{A}$), antenna structural scattering ($\vec{B}$), and load-dependent antenna mode scattering ($\vec{C}$).

Since the data rate of existing UHF RFID tags is typically much greater than the rate of tag physical motion or nearby object motion, and load-dependent vector $\vec{C}$ changes with the data rate, the contributions from environmental scattering and structural scattering ($\vec{A} + \vec{B}$) may be assumed constant during a particular symbol. The received field (10) then reduces to

$$E_{\text{rcv}} = E_{\text{env}} + E_{\text{scat}}(Z_L^*) + \Gamma^* \frac{E_{\text{env}I_m}}{I_{\text{ant}}} + N.$$ 

(12)

The stationary field component is mixed down to DC by the reader’s homodyne receiver and rejected by the receiver’s AC-coupling or DC offset compensation loop leaving only the desired modulated field containing data. Homodyne detection and DC rejection demodulates the load dependent scattering component $\vec{C}$ to yield the desired data.

The noise-free case is shown in Fig. 3(a). In this figure, a modified Smith chart normalized to $Z_0 = Z_0^*$ is positioned to show the range of the load-dependent antenna-mode scattering contribution ($\vec{C}$) when only passive load impedances are considered.

Movement of nearby scatterers results in a change in the environmental multipath contribution. This translates the tag’s total scattered field ($\vec{B} + \vec{C}$) within the I/Q plane as represented in Fig. 3(b). In contrast, as shown in Fig. 3(c), changing the tag-to-reader separation distance rotates total scattered field ($\vec{B} + \vec{C}$) due to the distance-dependent phase shift and scales it in amplitude because of changing path loss. In the multipath case, as well as the moving-tag case, there is a fixed phase relationship between the structural scattering and antenna-mode scattering which is desirable because it preserves the transmitted constellation.

This model has been confirmed experimentally through measurements described below.

VI. QAM TAG MODULATOR DESIGN AND SIMULATION

Based on the observation from (7) that a careful choice of modulating load impedance $Z_L$ can yield a scattered E-field component in any quadrant of the complex plane, a series of modeling exercises were conducted to simulate a practical backscatter QAM modulator. Component choices were validated using Agilent ADS with the simplified reader system model shown in Fig. 4.

Modulating impedance values can be chosen by first writing each symbol of the desired I/Q constellation in the form

$$S_1 = x_1 + jy_1$$

(13)

where $x_1$ represents the in-phase component and $y_1$ represents the quadrature component of the $i$-th symbol. In order to produce impedance values realizable with passive components, all reflection coefficients are confined within a circle about the conjugate match with magnitude $\leq 1$. These reflection coefficients are then scaled by a constant $0 < \alpha \leq 1$.

$$\Gamma_1^* = \alpha \cdot \frac{S_1}{\max |S|}$$

(14)

Values of $\alpha$ closer to 1 reflect increasing amounts of the incident RF power back to the reader and thus result in higher backscatter signal strength. This is the typical case for a semi-passive (battery-assisted) tag, such as the one
incident field. Values of $\alpha \ll 1$ are typical for a passive tag to permit a majority of the incident field to be absorbed in the energy harvesting circuit. Because $\alpha$ is a constant that relates power reflection and power transmission coefficients, the optimal value will depend on the desired balance between backscattered signal power and power delivered to the energy harvesting circuit [11], [12]. For binary ASK and PSK, this tradeoff has previously been analyzed as a function of reflection coefficient [13]. Due to government regulations limiting the reader’s transmitted power, well-designed passive RFID systems usually operate in a forward-link limited regime where the read distance is limited by the ability to deliver power to the tag.

By rearranging the conjugate reflection coefficient $\Gamma^*$ from (4), we find a set of complex impedance values for a given I/Q constellation

$$Z_{L_1} = \left. \frac{Z_a^* - Z_o\Gamma_i^*}{1 + \Gamma_i^*} \right|_{Z_a^*}$$

(normalized to $Z_a^*$, the conjugate of the antenna impedance. The resulting modulating impedances will then fall in all 4 quadrants of a modified Smith chart normalized to $Z_0 = Z_a^*$.

Because UHF RFID antennas are typically designed to include significant amounts of inductance in order to nearly conjugate match a capacitive tag IC impedance, all QAM modulating impedance values may be realizable using only R and C components. For example, an antenna designed to match the Impinj Monza4 IC [14] has an impedance $Z_a = 11 + j143$. The required impedance states for a 4-QAM/4-PSK modulator matched to this antenna with $\alpha = 1$ are all capacitive $[-j147.6, -j169.6, -j116.4, -j138.4]$ with $C \approx [1.18 \text{ pF}, 1.03 \text{ pF}, 1.49 \text{ pF}, 1.26 \text{ pF}]$ respectively.
TABLE I

<table>
<thead>
<tr>
<th>Ideal Impedance States: ( Z_{a} = 50 \Omega )</th>
<th>( \Gamma ) Error versus Frequency (Referenced to 915 MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>( Z )</td>
<td>R + jX ( \Omega )</td>
</tr>
<tr>
<td>( Z_{1} )</td>
<td>0.0 + 20.71j</td>
</tr>
<tr>
<td>( Z_{2} )</td>
<td>0.0 - 120.71j</td>
</tr>
<tr>
<td>( Z_{3} )</td>
<td>0.0 + 120.71j</td>
</tr>
<tr>
<td>( Z_{4} )</td>
<td>0.0 + 20.71j</td>
</tr>
</tbody>
</table>

TABLE II

<table>
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<th>Ideal Impedance States: ( Z_{a} = 50 \Omega )</th>
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</tr>
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<tbody>
<tr>
<td>( Z )</td>
<td>R + jX ( \Omega )</td>
</tr>
<tr>
<td>( Z_{1} )</td>
<td>0.0 + 30.9j</td>
</tr>
<tr>
<td>( Z_{2} )</td>
<td>0.0 - 80.9j</td>
</tr>
<tr>
<td>( Z_{3} )</td>
<td>130.9 + 0.0j</td>
</tr>
<tr>
<td>( Z_{4} )</td>
<td>19.1 + 0.0j</td>
</tr>
</tbody>
</table>

However, for a resonant antenna with approximately real-valued \( Z_{a} \), modulating impedances found by solving (15) will require R, L, and C components. This creates a problem for monolithic implementation because standard CMOS processes yield good quality resistors and capacitors but physically large and low Q inductors. In order to make CMOS integration feasible by eliminating L components, a transformation from RLC impedances to RC-only may be found by enforcing the constraint that

\[
\Gamma_{\text{RC only}}^{*} = \beta \left( \Gamma_{i}^{*} - j \min \left[ \Im \left( \Gamma_{i}^{*} \right) \right] \right)
\]

for all \( \Gamma_{i}^{*} \) before re-applying the conversion to impedance values (15), where \( \Im \left( \Gamma_{i}^{*} \right) \) represents the imaginary portion of the reflection coefficients and

\[
\beta = \frac{\max \left[ \Im \left( Z_{L} \right) \right]}{\max \left[ \Im \left( Z_{L} \right) \right] - \min \left[ \Im \left( Z_{L} \right) \right]}
\]

scales the reflection coefficients by the portion of impedances above the real line. This transformation compresses the I/Q vector space of the constellation, resulting in a tradeoff of reduced backscatter amplitude compared to the RLC implementation but yielding the desired QAM constellation with a CMOS process. Smith charts showing the conjugate reflection coefficients of the RLC and RC-only constellations are shown in Fig. 5 for a \( Z_{a} = 50 + j0 \Omega \) antenna where the compression of vector space spanned by adjacent symbols is apparent.

A. Experimental Validation

Based on the block diagram in Fig. 6, a discrete component semi-passive tag pictured in Fig. 7 was constructed to test a 4-state PSK / 4-QAM backscatter modulator designed with this procedure. Four lumped impedances are connected to an antenna port through an Analog Devices ADG904 SP4T CMOS FET switch that is controlled by a TI MSP430F2011 microcontroller. The microcontroller was programmed to stream 128 bit strings of pseudo-random data to modulate the 4 switch states at 200k symbols/s, corresponding to a data rate of 400kbps. The static DC power consumption of the modulator, excluding the power consumption of the microcontroller, is 3.0 V at 38.3 nA, or 115 nW. The dynamic power consumption is shown in Fig. 8. In the prototype tag, a MSP430F2011 microcontroller consumes an additional 1-2mA (or 3-6mW) when generating the pseudo-random data. A CR2032 3V lithium coin cell battery serves as a power source for the device.

The values of the lumped impedances were designed using (16) to match an antenna with \( Z_{a} = 50 + j0 \Omega \) and \( \alpha = 1 \). The calculated values for the L&C and C-Only cases are shown in Tables I and II. At a design center frequency of 915 MHz, the nearest standard lumped values were 8.0 pF, 2.2 pF, 22 nH, and 3.9 nH for the L&C case and 5.6 pF, 2.2 pF, 130 \( \Omega \) and 19.1 \( \Omega \) for the C-Only case. The capacitors have a tolerance of ±0.1 pF and the inductors have a tolerance of ±0.1nH. As the carrier frequency deviates from the design center frequency, constellation distortion is introduced causing a greater error vector magnitude (EVM), reflecting increasing bit error rate (BER).

Because the prototype semi-passive tag is matched to 50 \( \Omega \), a cabled test bench setup similar to that shown in Fig. 4 can be used to characterize the tag. An Agilent MXG N5181A synthesized RF source set to produce +3 dBm at 915 MHz was used as the forward link source. The reflected signal
was separated using an MiniCircuits ZFDC-15-10 15 dB directional coupler and fed to a Linear Technology LT5575 I/Q demodulator. The baseband I and Q signals were sampled at 8 bit resolution with a sampling rate of 2 MSPS using an Agilent Infinium DSO8104 oscilloscope. Fig. 9(a) shows the received constellation generated by the L&C modulator while Fig. 9(b) shows the received constellation generated by the C-Only modulator as captured by the oscilloscope with random I and Q data.

Digitally generated AWGN was summed with the captured time-domain signal samples. A minimum distance soft decision demodulator was used to obtain BER curves for the two constellations (L&C, and C-Only) found in Tables I and II and the Smith charts of Fig. 5. The resulting measured BER curves, shown in Fig. 10, showed good agreement with those predicted by MATLAB simulation.

The final step was to connect the switched impedances to a 50 Ω dipole antenna and perform an over-the-air test of the resulting semi-passive backscatter modulator. The transmitter employed for this test consisted of an Agilent MXG N5181A synthesized signal source followed by a Mini-Circuits ZHL-5W-2G amplifier yielding a conducted output power of +30 dBm. The EIRP of this transmitter setup was +38.4 dBm given a Yagi antenna of 9dBi gain and a cable loss of 0.6 dB. A 6 dBi patch receiving antenna was located with close proximity to the transmitting antenna to minimize the radar bistatic angle. The test setup used for gathering the over-the-air data is shown in Fig. 11.

As the tag moved radially outward with respect to the reader, the baseband constellation was observed to rotate and scale as expected given the signal model presented in Fig. 3. Due to the indoor environment and the interaction with multipath signals, some locations in the lab exhibited a deep fade in the detected backscatter signal. This behavior is typical for any passive or semi-passive backscatter system and is not specific to the QAM modulation method presented in this work.

During initial testing in a typical indoor lab environment, 500,000 bits were transmitted from multiple locations. No more than 1 bit error was observed at any location within 4.5 m. This implies a BER of less than $10^{-5}$ over the air in a low-interference, relatively controlled laboratory setting. Given the experimental setup, measurement times and thus the number of bits accumulated for error rate measurements are limited by oscilloscope capture memory depth. The tag was then fixed at a separation of 2.92 m from the transmit antenna and the RF power feeding the antenna was swept to produce
the data shown in Fig. 12. Simulation as well as over-the-air results show the C-Only 4-QAM / 4-PSK constellation performance trailed the L&C performance by a margin of approximately +1.5 dB Eb/N0 to achieve equivalent BER. This is a tradeoff due to the reduction in inter-symbol distance brought about by the compression of the C-Only constellation by the factor $\beta$.

VII. DISCUSSION

The development of passive and semi-passive RFID system models has in most cases followed directly from radar systems techniques. While scalar valued differential radar cross section (DRCS) analysis is useful for analyzing binary ASK or PSK modulation schemes, the DRCS approach is unnecessarily limiting. Because the $E$ field scattered from a tag antenna is complex-valued, passive and semi-passive RFID systems can be designed to produce $M$-ary QAM signals with minimal increase in cost or complexity at the tag or the reader.

More versatile $M$-ary modulation schemes also permit an important new design space for RFID systems. Because a tag’s on-chip oscillator frequency ultimately limits the symbol rate, but not the bit rate, higher values of $M$ yield increased data rate for a given on-chip oscillator frequency. This in turn allows for fine grained tradeoffs between on-chip power consumption, data rate, and link margin.

In terms of bandwidth (spectral) efficiency and performance in the presence of AWGN, 4-QAM/4-PSK and binary PAM have equivalent performance [15]. For a given data rate, the lower symbol rate of 4-QAM/4-PSK permits the QAM modulator transistors to switch at half the frequency compared to a binary modulator. In general, an $M$-ary QAM modulator can be clocked slower by a factor of $\log_2 M$. In the digital sections of a tag IC, careful choice of internal memory bus width and associated data path width may also allow the clock rate driving these on-chip circuits to be correspondingly reduced. The power dissipation in all of these sections will thus decrease in accordance with the $\frac{1}{2}CV^2f$ model for CMOS power dissipation. Higher order constellations can improve modulator power efficiency further.

As an example, a 16-QAM constellation is shown in Fig. 13 for implementation either with or without inductors. When designed for a 50 $\Omega$ antenna operating at 915 MHz, the $\alpha = 1$ C-Only implementation requires a minimum capacitance of 1.93 pF, a maximum capacitance of 12.74 pF, and a maximum resistance of 130.9 $\Omega$. Future work will explore the tradeoffs in the choice of $\alpha$ and $\beta$ with such higher-order constellations.

The multi-state RF switch based modulator presented in this paper is simple and power efficient, though it trades off die area for simplicity because $M$ switched impedances are required to implement $M$-ary modulation. Future modulator designs can be realized with weighted combinations of $R$, $L$, and $C$ components to form simple impedance DACs that hold the promise of similar power consumption with reduced die area.

Finally, as the behavior of backscatter links in multipath environments becomes better understood, it is expected that independent control of the backscatter amplitude and phase will allow for modulation schemes to be designed with multipath immunity as a primary goal.

VIII. CONCLUSION

Device-level simulation and measurements of a 4-PSK / 4-QAM modulator are presented for a semi-passive (battery-assisted) tag operating in the 850-950 MHz band. This first prototype modulator consumes a static power of only 115 nW but is capable of transmitting 4-PSK / 4-QAM data at a rate of 200k symbols/s or 400 kbps. The measured power consumption shows a linear increase in modulator power as the data rate increases, as expected from the $\frac{1}{2}CV^2f$ power dissipation model for CMOS switching devices. While power consumption increased with the symbol rate due to the switching devices used in the prototype, it remained insignificant when compared to the consumption levels of the low-power Texas Instruments MSP430 microcontroller.

When compared with binary ASK or PSK based RFID systems, tags employing vector backscatter modulation hold the promise of higher data throughput while running at a lower on-chip clock frequency, thereby reducing on-chip power consumption and extending read range. Alternatively, tags employing an $M$-ary modulator can achieve $\log_2 M$ higher data throughput at essentially the same DC power consumption as a tag employing binary ASK or PSK.
Data in this paper was obtained using a discrete component semi-passive tag, but the methods are equally applicable to fully passive tags. Future work will expand upon this system by implementing higher order constellations, improvements to signal processing for multipath robustness, and finally implementing fully passive tags. Future work will expand upon this system and

Fig. 13. Ideal impedances for a $M = 16$ QAM modulator with $Z_a = 50\Omega$ and $\alpha = 1$, with and without inductors

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REFERENCES


