

QAM Backscatter for Passive UHF RFID Tags

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Abstract—Traditional passive UHF RFID tags employ either ASK or PSK backscatter modulation to communicate data from memory or sensors on the tag to a remotely-located reader. These simple modulation schemes transfer data at a rate of one bit per symbol period, which for an integrated CMOS tag IC requires an on-chip oscillator with a frequency at least equal to the bit rate. Motivated by the fact that most modern UHF RFID readers already employ I/Q demodulation of the backscattered signal to account for backscatter phase rotation as the tag moves with respect to the reader, we propose a QAM backscatter method using no on-chip inductors that is compatible with a single-chip CMOS tag implementation. With QAM backscatter, tags transmit more than one data bit per symbol period, permitting tag designers to employ a lower power on-chip oscillator operating at a frequency equal to the (lower) symbol rate while maintaining the same data throughput as ASK or PSK, or alternatively to send data at a higher rate for a given on-chip oscillator frequency. We present the fundamental design equations required for arbitrary QAM backscatter modulators and present simulated bit error rate (BER) and error vector magnitude (EVM) curves for the operation of an inductor-free 4-QAM and 8-QAM modulator centered at 915MHz and evaluated over the 860MHz-950MHz worldwide UHF band.

I. INTRODUCTION

Radio frequency identification (RFID) technology was initially conceived around the notion of transferring only a unique ID (UID) or a small amount of on-chip memory from a tag to a reader. In particular, low-cost passive UHF “smart label” transponders, including those based on EPC Global’s Generation 2 specification [1], need to derive all tag operating power from an incoming reader RF carrier. This results in a drive toward IC simplicity and low power CMOS design techniques. Minimizing tag IC complexity and on-chip clock frequencies results in lower tag operating power, which is crucial for maximizing read range in a power-limited passive UHF RFID system. At the same time, simplifying the tag design results in a smaller tag die, more tag die on a given wafer and improved yield, and thus lower overall tag cost.

As passive UHF RFID technology has matured, many new application scenarios have been proposed where a tag is expected to transmit ever-increasing amounts of data. These scenarios include tags with expanded on-chip memory of 128KB or more, tags including complex cryptographic security protocols, or tags that transfer stored sensor data in a semi-passive mode. An extreme example of this trend is the Intel WiSP, a passive UHF RFID platform including a fully accessible, programmable 16-bit microcontroller with a variety of sensor peripherals [2]. The WiSP platform is

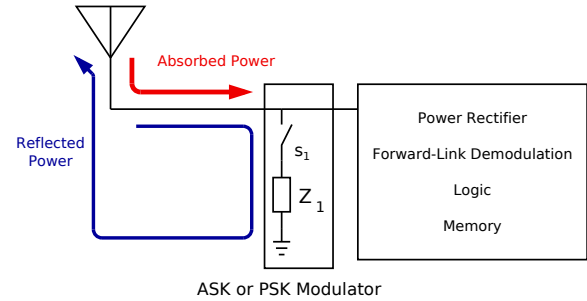


Fig. 1. Simplified UHF RFID tag model with single ASK / PSK modulating switch.

currently being used for a variety of research applications [3]–[6]. As RFID tags include more data intensive sensors and more computational power, the achievable data rate vs. tag operating power of the existing ASK or PSK communication channel becomes increasingly limiting.

In most passive UHF RFID systems, reader-to-tag communication (the “forward link”) is an ASK or PR-ASK channel implemented by modulating either the amplitude, or both the amplitude and phase, of the reader’s transmitted RF carrier. A simple, low power envelope detector on the tag is used to recover forward link data. To transmit data from the tag to the reader (the “return link”), the tag modulates its radar cross section σ_{tag} , thus changing the amplitude or phase of its reflection. This is performed by modulating the load impedance of the tag die Z_{tag} as seen by the tag’s antenna as shown in Figure 1. In a perfectly matched condition, when the impedance presented by the tag IC is the complex conjugate of the tag antenna’s impedance, the tag’s absorbed power is maximized and reflected power back to the reader is minimized. If there is an impedance mismatch between the tag’s antenna and the tag IC, an additional portion of the received power is reflected and re-radiated by the antenna. By modulating the impedance across its terminals, the tag introduces a controlled mismatch impedance and thus creates a time-varying backscatter signal that is received by the reader.

This modulated backscatter communication link consumes much less power than an active transmitter requiring an on-chip oscillator running at the RF carrier frequency. In a modulated backscatter return link, the on-chip clock oscillator runs at a multiple of the data link frequency (e.g. up to 1.28MHz) rather than a UHF carrier frequency in the 860-950MHz range. The return link modulator is also typically

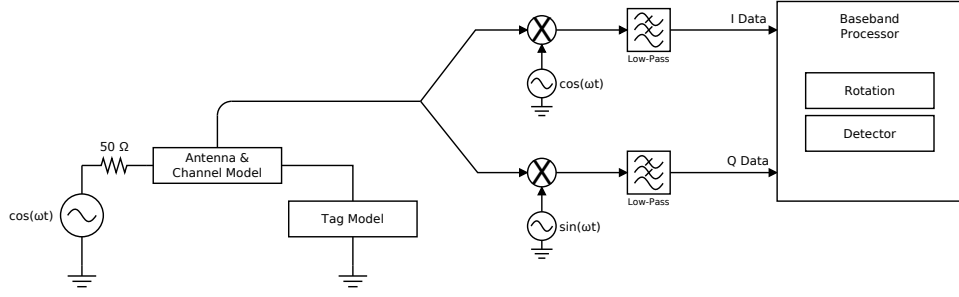


Fig. 2. The simplified reader model used in MATLAB and ADS simulations of the QAM modulator.

simplified to a switching FET that introduces two impedances across the tag's antenna terminals, and thus two modulation states for the tag's backscattered energy [7], [8].

Referring to Figure 1, ASK backscatter modulation is achieved by selecting a pure resistance for Z_1 , thus moving the real part of the tag IC's impedance between a matched state, which maximizes rectified DC power available to the tag IC, and a reflective state. PSK backscatter is performed by switching a load reactance, usually a capacitance, at Z_1 . PSK load reactances are typically capacitive as inductors are larger and more lossy than capacitors in most CMOS processes. While PSK can be more difficult to integrate with an antenna than ASK, it yields a lower bit error rate (BER) for a given signal to noise ratio when compared to an ASK scheme [9].

Most UHF RFID standards, including the EPC Global Generation 2 specification, permit tags to implement either ASK or PSK backscatter. Furthermore, tag backscatter phase with respect to the reader's local oscillator (LO) varies with distance from the reader to the tag, so received ASK and PSK backscatter constellations are rotated by an unknown angle with respect to the real axis. To achieve acceptable demodulator performance, most modern UHF RFID readers are based on the homodyne receiver concept with an RF I/Q demodulator as shown in Figure 2 followed by a baseband phase rotation block that rotates the incoming signal back to the in-phase axis prior to making a 1/0 bit decision. Existing UHF RFID reader hardware is therefore adequate to demodulate M-ary QAM data, but reader demodulation firmware would need to be rewritten to demodulate M-ary QAM regardless of the distance-induced phase rotation.

The proposed M-ary QAM backscatter technique employs $M + 1$ FET switches $S_1 \dots S_{M+1}$ along with M load impedances $Z_1 \dots Z_M$ to implement M-ary QAM modulation. Switch S_{M+1} is a normally-closed FET switch that allows the tag's power rectifier and forward link demodulator to operate in a conjugate-matched condition, while the remaining FET switches introduce M different load impedances to represent M different symbols in an M-ary QAM constellation. The proposed technique is general and can be implemented with or without inductors, but the most interesting results shown below are implemented without inductors and are therefore suitable for single-chip CMOS integration.

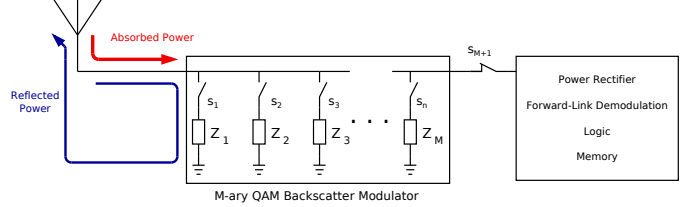


Fig. 3. Proposed tag modulator scheme showing M different load impedances $Z_1 \dots Z_M$

II. SELECTING LOAD IMPEDANCES $Z_1 \dots Z_M$

Each impedance in Figure 3 represents a single symbol. The magnitude and phase of the electric field reflected by a tag antenna with input impedance Z_a is determined by the antenna reflection coefficient Γ_a of the tag [10]. The scattered field is a function of $1 - \Gamma_a$ where $\Gamma_a = \frac{Z_a - Z_L}{Z_a + Z_L}$ and Z_L is the load impedance of the network attached to the antenna. By specifying the desired reflection coefficients $\Gamma_1 \dots \Gamma_M$, we can determine the set of impedances $Z_1 \dots Z_M$ for the M modulation states. To find these impedances, each symbol state in the desired QAM constellation is written in the form

$$(I_i, Q_i) \quad (1)$$

where

$$\begin{aligned} |I_i| &\leq 1 \\ |Q_i| &\leq 1 \end{aligned} \quad (2)$$

and each value I_i and Q_i represents the inphase and quadrature component of an ideal symbol in the constellation. For example, a possible 4-QAM constellation places the symbol locations at $(I, Q) = (1, 1), (-1, 1), (-1, -1), (1, -1)$. For modulation schemes with more symbols, the I/Q location spacing must be scaled per (2).

The I/Q values are then converted to complex reflection coefficients by

$$\Gamma_i = I_i + jQ_i. \quad (3)$$

For a 1-port network, we can then find the load impedance Z_i

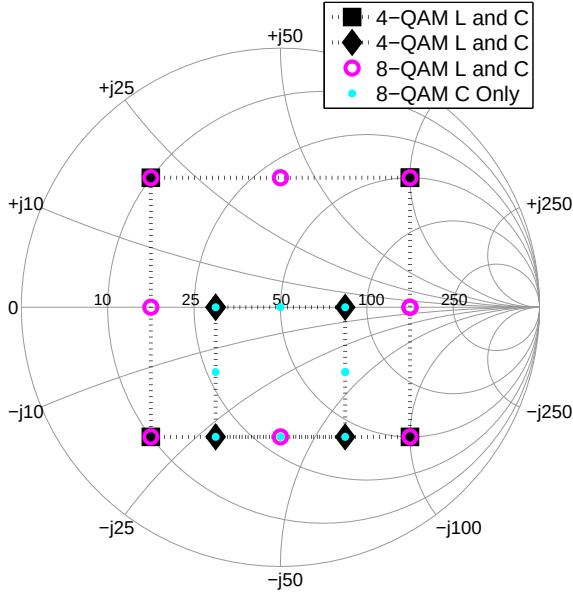


Fig. 4. Impedance states for 4-QAM (L & C), 4-QAM (C Only), 8-QAM (L & C) and 8-QAM (C Only) on a 50Ω Smith chart ($Z_a = Z_0 = 50\Omega$).

corresponding to each Γ_i :

$$Z_i = -Z_a \frac{\alpha \Gamma_i + 1}{\alpha \Gamma_i - 1} \quad (4)$$

$$\alpha < \frac{1}{\sqrt{2}}$$

The scale factor α must be selected to avoid a pole at $\Gamma_i = \pm 1$ and negative resistance values when $\alpha > \frac{1}{\sqrt{2}}$. As α approaches zero, the density of the impedance states near the antenna resonant impedance Z_a increases, which results in less power being reflected or re-radiated from the tag and thus a lower signal-to-noise ratio for the backscatter channel.

The Smith chart presented in Figure 4 shows the selected impedances $Z_{1...M}$ for three different QAM modulation schemes. Two of the constellations are implemented with both inductors and capacitors (L&C), while one of the 4-QAM constellations is constrained to only capacitors (C Only). The (C Only) constraint makes CMOS integration feasible and is achieved by finding

$$\Gamma_{iC \text{ Only}} = \frac{\Gamma_i}{2} - j \frac{\max \left\{ \left| \Im(\Gamma_i) \right| \right\}}{2} \quad (5)$$

for all Γ_i where $\Im(\Gamma_i)$ represents the imaginary portion. This compresses the I/Q vector space and translates all reflection coefficients such that (4) yields a set of impedances with reactances on or below the real line. The effect is visible in Figure 4 where the choices of impedances are normalized to $Z_a = Z_0 = 50\Omega$ and presented in Tables I, II, and III.

Each reactance value $Z_{1...Z_M}$ must be designed around a given antenna impedance Z_a and center frequency. Assuming an implementation centered at 915MHz and an antenna of

TABLE I
IMPEDANCE VALUES FOR 4-QAM (C ONLY, $\alpha = 0.5$)

Ideal Impedance States: $Z _{Z_0} = 50 \Omega$				Γ Error versus Frequency (Referenced to 915 MHz)	
Z	R + jX Ω	C Value	Γ	860 MHz	950 MHz
Z_1	30	-	-0.25	0%	0%
Z_2	83.33	-	0.25	0%	0%
Z_3	18.97 - j27.59	6.30 pF	-0.25 - j0.5	1.82%	1.16%
Z_4	42.3 - j61.54	2.83 pF	0.25 - j0.5	4.18%	2.54%

TABLE II
IMPEDANCE VALUES FOR 4-QAM (L&C, $\alpha = 0.5$)

Ideal Impedance States: $Z _{Z_0} = 50 \Omega$				Γ Error versus Frequency (Referenced to 915 MHz)	
Z	R + jX Ω	L/C Value	Γ	860 MHz	950 MHz
Z_1	50 + j100	17.39 nH	0.5 + j0.5	3.15%	1.86%
Z_2	10 + j20	3.48 nH	-0.5 + j0.5	0.59%	0.39%
Z_3	10 - j20	8.70 pF	-0.5 - j0.5	0.59%	0.39%
Z_4	10 - j100	1.74 pF	0.5 - j0.5	0.81%	0.47%

TABLE III
IMPEDANCE VALUES FOR 8QAM (L&C, $\alpha = 0.5$)

Ideal Impedance States: $Z _{Z_0} = 50 \Omega$				Γ Error versus Frequency (Referenced to 915 MHz)	
Z	R + jX Ω	L/C Value	Γ	860 MHz	950 MHz
Z_1	150	-	0.5	0%	0%
Z_2	16.667	-	-0.5	0%	0%
Z_3	30 + j40	6.96 nH	0 + j0.5	3.65%	2.28%
Z_4	30 - j40	4.35 pF	0 - j0.5	3.65%	2.28%
Z_5	50 + j100	17.39 nH	0.5 + j0.5	3.15%	1.86%
Z_6	10 + j20	3.48 nH	-0.5 + j0.5	0.59%	0.39%
Z_7	50 - j100	1.74 pF	0.5 - j0.5	3.15%	1.86%
Z_8	10 - j20	8.70 pF	-0.5 - j0.5	0.59%	0.39%

constant impedance, the error in achieved Γ versus frequency varies from zero for purely resistive states to $\approx 4\%$, resulting in increasing degradation in constellation accuracy as the operating frequency is varied from the design center frequency. The effect of frequency induced impedance error in terms of constellation accuracy is shown in Figure 5 while the impact on BER vs. E_b/N_0 is presented in Figures 7 and 6.

III. SIMULATION

The proposed QAM modulation schemes were evaluated using a series of MATLAB and Agilent ADS models based on the reader model of Figure 2. Various levels of additive white Gaussian noise (AWGN) were added to evaluate bit error rate (BER) versus E_b/N_0 . Several exemplary I/Q constellations including the effects of AWGN as well as variation with frequency, given reactances designed for a 915MHz center frequency, are shown in Figure 5.

As expected, for a constant $BER \approx 10^{-3}$, the 8-QAM (L&C) scheme has about 3dB of E_b/N_0 penalty versus 4-QAM (L&C), which has the best BER performance. The ≈ 2 dB performance penalty of the 4-QAM (C Only) constellation is due to the compression of the I/Q vector space, and the corresponding reduction of tag differential RCS, $\Delta\sigma$, between any two symbols. The error vector magnitude (EVM) for the modulation schemes is shown in Fig 7, where EVM is defined as in the IEEE 802.11 specification [11].

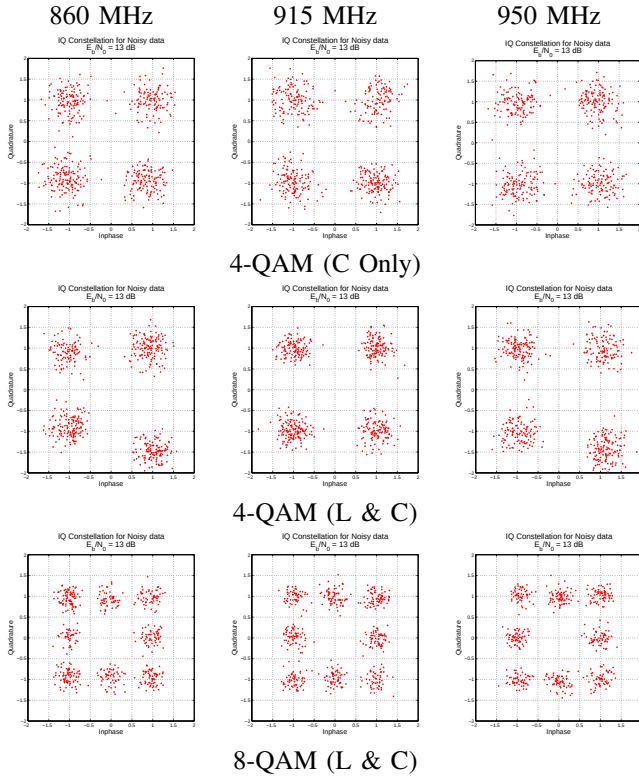


Fig. 5. I/Q Constellations for 4-QAM and 8-QAM, $E_b/N_0 = 13\text{dB}$

$$\text{EVM} = 100 \cdot \sqrt{\frac{\frac{1}{N} \sum_{k=1}^N |\hat{S}_{k\text{Ideal}} - S_{k\text{Sample}}|^2}{\frac{1}{N} \sum_{k=1}^N |\hat{S}_{k\text{Ideal}}|^2}} \quad (6)$$

where N is the total number of symbols transmitted, $S_{k\text{Sample}}$ is the normalized k^{th} demodulated symbol location in the I/Q plane, and $\hat{S}_{k\text{Ideal}}$ is the ideal (noiseless) normalized symbol position. The EVM represents the average percentage difference of the symbol location error and the magnitude of the ideal symbol location.

A. ADS Device-Level Simulation

To further validate the performance of the QAM modulator scheme, a device-level simulation was performed using the Agilent Advanced Design System (ADS). The reactances for the 4-QAM (C Only) were realized using resistors and capacitors of finite Q . In this way, all four states were modeled as real components using the standard ADS libraries.

As before, AWGN was added to the signal and the results were analyzed. The bit error rate versus bit energy is shown in Fig 6. The device level simulations show good agreement in bit error rate versus E_b/N_0 when compared with the predictions of the MATLAB model.

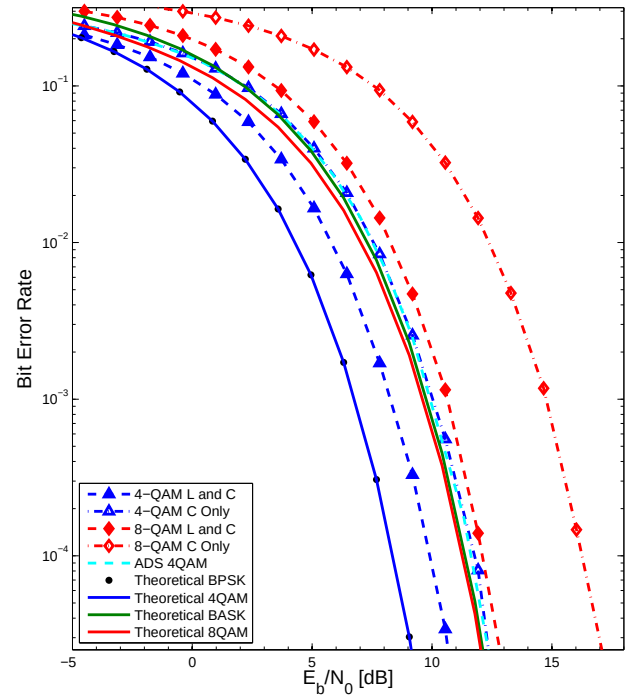


Fig. 6. MATLAB simulation of BER versus E_b/N_0 for 4-QAM and 8-QAM constellations implemented with and without inductors. ADS device level simulation data is shown for the 4-QAM C Only case. The theoretical limits corresponding to ideally-implemented binary ASK, PSK, and QAM are shown for comparison.

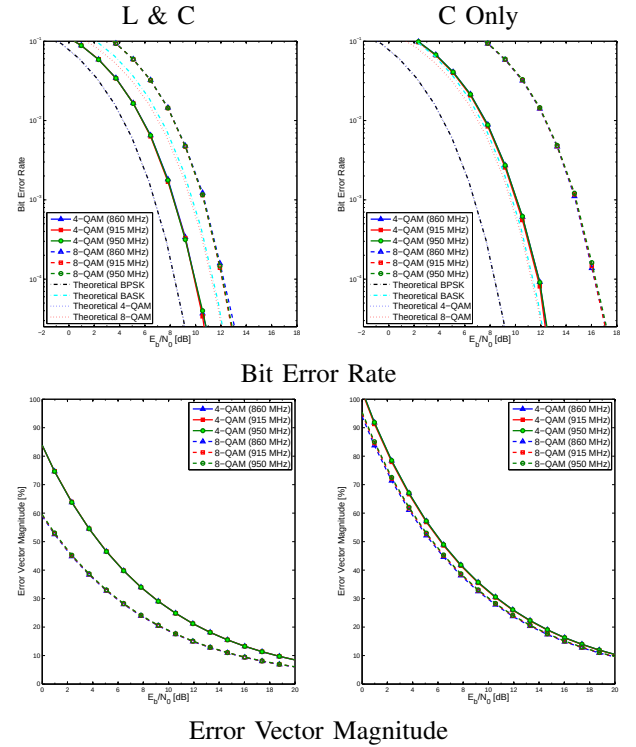


Fig. 7. BER and EVM versus frequency over 860MHz-950MHz for design center frequency of 915MHz.

IV. DISCUSSION AND FUTURE WORK

In this paper we have presented a methodology for designing an M-state QAM backscatter modulator that is suitable for single-chip CMOS implementation. For 4-QAM and 8-QAM constellations, our MATLAB and ADS device-level simulations indicate that acceptable BER performance of 10^{-3} with E_b/N_0 between 9dB and 11dB can be obtained from passive UHF tags operating over the entire 860MHz - 950MHz band given only a single set of modulating impedances designed for a 915MHz center frequency. These results are several dB worse than the textbook results for coherent QAM demodulation due to the restriction of allowable symbol placement in the I/Q plane when only physically realizable component values are allowed. BER performance is further reduced by the addition of the scalar α and the (C Only) mapping. In general, well designed passive UHF RFID systems are forward-link limited and there is normally excess return-link margin at the tag's power-up limit. This suggests that there is an opportunity to trade off excess SNR for improved data rate, even given the BER penalties shown in this work. The extent of this design tradeoff between BER performance and realizable circuits will be explored in future work.

Additional future work includes constructing a discrete component semi-passive tag emulator that implements QAM modulation and evaluating real-world performance in the presence of multipath in indoor and outdoor scenarios. We will also improve our ADS model to further prove the viability of CMOS integration by including extracted parameters from test devices actually implemented in various CMOS processes.

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