Baseband Signals and Power in Load-Modulated Digital Backscatter

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***Abstract*—Passive digital backscatter signals in systems like radio frequency identification (RFID) are usually received along with strong interference from a leaked carrier. The simplest way to quantify the "useful" communication signal is to separate it as an amplitude-shift keying (ASK) or biphase-shift keying (BPSK) component. These definitions give different power normalizations, posing some complexity in comparison of link quantities. This letter investigates their suitability in terms of basic signal theory and conservation of energy to clarify relationships between the baseband signals and "backscattered power." Defining received backscatter as BPSK guarantees energy conservation for arbitrary tag modulation loads.**

***Index Terms*—Electromagnetic scattering, modulation, radio communication, radio frequency identification (RFID).**

I. Introduction

THE EXPANDING use of radio frequency identification (RFID) over the past decade has driven new study in backscatter communication. Today, it is still less well understood than powered transmission. Basic parameter definitions remain unclear, clouding the meaning of theory and data.

Simple backscatter communication is compared against transmission in Fig. 1. The tag chip modulates reflected waves by switching the impedance connected to its antenna. The reader effectively broadcasts the local oscillator (LO), making an on-chip RF phase-locked loop (PLL) or oscillator unnecessary. This reduces tag chip area and power consumption, but reflects weak power to readers, limiting tag-to-reader (return) link range and increasing reader complexity. Fully passive tags are simpler still, replacing battery power with rectified LO, but limiting reader-to-tag (forward) link range.

Current standards [1], [2] permit tags to backscatter either amplitude-shift keying (ASK) or biphase-shift keying (BPSK) modulation, but define neither. Most prior work seems to interpret these as "ASK loads" and "BPSK loads," referring to the complex power-wave representation of the modulator states [3]–[5]. Separately, the modulation signal received by a reader can also be defined as ASK or BPSK voltages, but these give different power normalizations. As a result, a signal in [6] and [7] has twice the "backscattered power" as the same signal in [4] and [8] (or half of that in [5]).

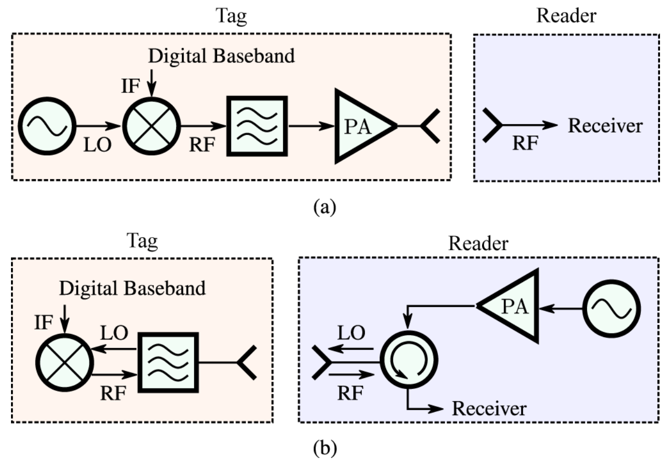

Fig. 1. Comparison between simple (a) active (transmitting; ASK) against (b) passive (backscattering; load-modulated) communication topologies.

This letter discusses the underlying assumptions and side effects of ASK or BPSK power normalizations, focused on application to passive UHF RFID. We first consider separating the backscatter and leaked interference baseband components and encoding-specific effects of the frequency-modulated backscatter from tags. These lead to expressions for ASK- and BPSK-defined normalizations in terms of baseband signals. Finally, we compare how each normalization impacts fulfillment of the passivity condition wherever "received power" is defined.

II. Signal Definitions

Consider a received real-valued time-domain voltage

$$v(t) = \sqrt{2}(|V_{\text{leak}}|\cos(2\pi f_{\text{c}}t + \angle V_{\text{leak}}) + |V_{\text{bs}}(t)|\cos(2\pi f_{\text{c}}t + \angle V_{\text{bs}}(t)). \quad (1)$$

The two amplitude terms are the RMS leaked transmitter carrier and tag modulation voltages, and f_{c} is carrier frequency.

Bedrodian's product theorem [9] lets us convert this narrowband modulated signal to a Gabor analytic signal [10], $\mathcal{V}$

$$\mathcal{V}(t) = \sqrt{2}e^{j2\pi f_{\text{c}}t}(V_{\text{leak}} + V_{\text{bs}}(t)). \quad (2)$$

It is related to the time-domain signal as $v(t) = \text{Re}(\mathcal{V})$. The complex RMS baseband V_{leak} and $V_{\text{bs}}(t)$ are as in (1). Their sum, $V(t) = V_{\text{leak}} + V_{\text{bs}}(t)$, is the baseband in-phase and quadrature (IQ) output from an ideal lossless demodulator.

We follow here the Fourier transform defined as

$$\mathscr{F}[v](f) = \int_{-\infty}^{+\infty} v(t)e^{-j2\pi ft}dt \quad (3)$$

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as defined by many instrument manufacturers in terms of unitary frequency rather than radial frequency [11]–[13]. This is related to the positive half-space of the transformed analytic signal as [14, p. 9]

$$\mathscr{F}[v](f) = \begin{cases} \frac{1}{2}\mathscr{F}[\mathcal{V}(t)](f), & f > 0 \\ \mathscr{F}[\mathcal{V}(t)](0), & f = 0 \\ \frac{1}{2}\mathscr{F}[\mathcal{V}(t)](-f) & f < 0. \end{cases} \tag{4}$$

Placing (2) into (4) and assuming no dc component in $v(t)$ gives

$$\mathscr{F}[v](f) = \begin{cases} \frac{\mathscr{F}[e^{j2\pi f_c t}(V_{\text{leak}}+V_{\text{bs}}(t))](f)}{\sqrt{2}}, & f > 0 \\ 0, & f = 0 \\ \frac{\mathscr{F}[e^{j2\pi f_c t}(V_{\text{leak}}+V_{\text{bs}}(t))](-f)}{\sqrt{2}}, & f < 0. \end{cases} \tag{5}$$

Ideal mixing with f_c, represented by the pairs $\mathscr{F}[f(t)e^{j2\pi f_0 t}](f) = \mathscr{F}[f(t)](f - f_c)$ and $\mathscr{F}[e^{j2\pi f_c t}](f) = \delta(f - f_c)$, leads to

$$\mathscr{F}[v](f) = \frac{1}{\sqrt{2}}(V_{\text{leak}}[\delta(f+f_c) + \delta(f - f_c)] + \mathscr{F}[V_{\text{bs}}](f - f_c) + \mathscr{F}[V_{\text{bs}}](f + f_c)). \tag{6}$$

Spectrum analyzers often show power spectral density (PSD) defined only in the positive half-space of the frequency domain. This includes power from negative frequency components "folded" onto the positive half space

$$\begin{aligned} \text{PSD}[v(t)](f) &= \frac{2|\mathscr{F}[v](f)|^2}{\text{Re}(Z)}, \qquad f \geq 0 \\ &= \begin{cases} \frac{|V_{\text{leak}}+\mathscr{F}[V_{\text{bs}}](0)|^2}{\text{Re}(Z)}, & f = f_c \\ \frac{|\mathscr{F}[V_{\text{bs}}](f-f_c)|^2}{\text{Re}(Z)}, & f > f_c. \end{cases} \end{aligned} \tag{7}$$

This is the "ideal" continuous PSD, neglecting realistic measurement effects like discretization, compression, uneven frequency response, spurious harmonics, impedance mismatch, and windowing. It is defined for $f \geq 0$ and normalized to the real part of the instrument port impedance, $\text{Re}(Z)$.

III. Received Backscatter Signaling

A. Separating Modulation and Leakage

"Backscattered power" depends on $V_{\text{bs}}(t)$, so we need to separate detected V into a useful communication component $V_{\text{bs}}(t)$ and interfering leaked component V_{leak}. Defining the received modulation (not the loads inside the tag) as purely ASK or BPSK results in the definitions illustrated in Fig. 2, with the two discrete detector states V_1 and V_2.

In the "offset ASK" case of Fig. 2(a), let "leaked" offset state be $V_{\text{leak}} = V_2$. This makes the decomposition

$$\begin{aligned} V_{\text{leak}}\ (\text{ASK case}) &= V_2 \\ V_{\text{bs}}\ (\text{ASK case}) &= \{0, V_1 - V_2\} = \{0, \Delta V\}. \end{aligned} \tag{8}$$

The interfering V_2 and modulation ΔV are independent of each other for the typical case of large leaked V_2.

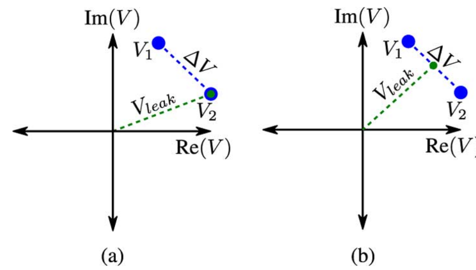

Fig. 2. Digitally modulated baseband backscatter signal can be decomposed into $V(t) = V_{\text{bs}}(t) + V_{\text{leak}}$ as (a) offset ASK (ASK + leaked offset) or (b) offset PSK (BPSk + leaked offset).

Now consider the "offset BPSK" of Fig. 2(b)

$$\begin{aligned} V_{\text{leak}}\ (\text{BPSK case}) &= \frac{1}{2}(V_1 + V_2) = \overline{V} \\ V_{\text{bs}}\ (\text{BPSK case}) &= \pm\frac{1}{2}(V_1 - V_2) = \pm\frac{1}{2}\Delta V. \end{aligned} \tag{9}$$

This time, baseband $\overline{V}$ and $\Delta V/2$ are the interfering and modulation components to characterize BPSK $V(t)$.

B. Passive RFID Backscatter Modulation Encoding

The symbol encoding affects the dependence of the leaked component on the symbol values. By (7), the ASK $V_{\text{leak}} = V_2$ or BPSK $V_{\text{leak}} = \overline{V}$ cases represent only carrier power. The carrier component depends on the dc component of $\mathscr{F}[V_{\text{bs}}]$, which in turn depends on the data encoding.

Passive UHF RFID symbol encoding defined by standards is biphase frequency modulation [1], [2]. This already hints that BPSK may be a natural definition of V_{mod}. Standards call the faster signal switching rate, f_m, the "link frequency."

Signal polarity switches at the beginning of each binary symbol. These symbols are differentiated by an extra polarity switch during each $1/f_m$ period, making this a frequency modulation scheme. The symbol rate is f_m/M [2], where the signal parameter M is the "number of subcarrier cycles per symbol." Encoding with $M = 1$ is called "FM0," and $M = \{2, 4, 8\}$ are "Miller." The FM0 and Miller encoding polarities are opposite: FM0 switches fastest on binary "1" symbols, and Miller switches fastest on binary "0." On average, these schemes should occupy each state of $V_{\text{bs}}(t)$ with about 50% duty cycle.

Fig. 3 shows the baseband time-domain and numerically computed RF spectrum of ASK $v_{\text{bs}}(t)$ for a train of FM0-encoded hexadecimal values `FFFFFFFF`... (a rectangular pulse train) and the random 32-bit value `DEADBEEF`. The link rate is the maximum allowed by standards [1], [2]. The "slurred" sidebands in the `DEADBEEF` signal are as expected for the combination of frequency-modulated binary "0" and "1" symbols. At 50% duty cycle between states $\{0, \Delta V\}$, the spectrum shows the expected dc component of $\Delta V/2$.

Fig. 4 shows the same signal as Fig. 3 interpreted as BPSK. Baseband now switches between $\pm\Delta V/2$. The sidebands are unchanged, but at 50% duty cycle the carrier component is gone.

Is zero-mean $V_{\text{bs}}(t)$ reasonable to expect in practice? A potential source of baseband dc is an odd number of "long"

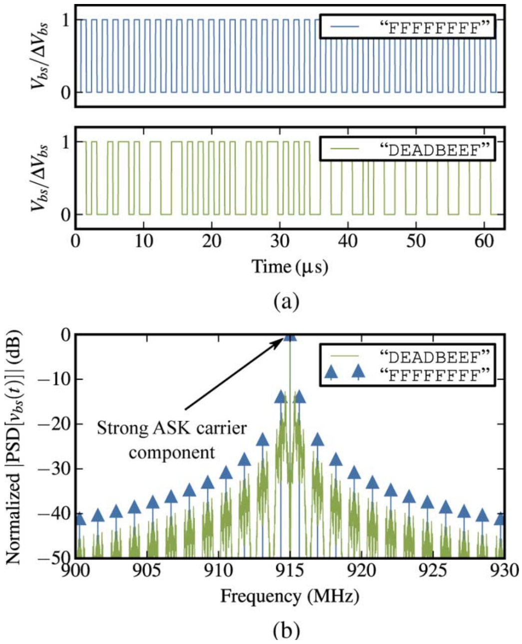

Fig. 3. Modulation component of repeated ASK FM0-encoded hexadecimal values `FFFFFFFF` and `DEADBEEF` in the (a) time and (b) frequency domains. The link rate is $f_m = 640$ kHz.

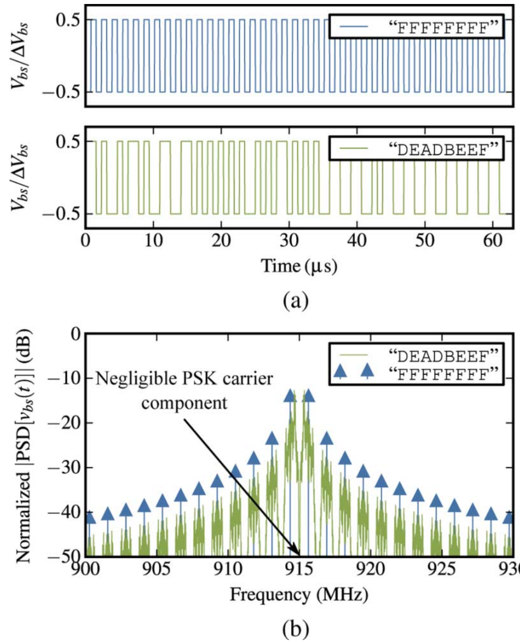

Fig. 4. Modulation component of repeated BPSK FM0-encoded hexadecimal values `FFFFFFFF` and `DEADBEEF` in the (a) time and (b) frequency domains. The link rate is $f_m = 640$ kHz.

modulation states: FM0-encoded data "0," or Miller-encoded data "1." This error is

$$|\max[V_{\rm bs}\text{DC offset}]| = \frac{\left|\frac{\Delta V}{2}\right|}{M \times (\text{Number of bits in } [T_1, T_2])}. \tag{10}$$

A data stream containing a 96-bit tag electronic product code (EPC) identification number has error below $0.01|\Delta V/2|$ for the FM0 $(M = 1)$ case, 40 dB smaller than the modulation power. This error is even smaller for Miller modulation $(M = \{2, 4, 8\})$.

Another source of baseband dc could be uneven time in each state. Current standards require duty cycle between 45%–55%. This dc offset in BPSK $V_{\rm bs}(t)$ from a "standard-compliant" tag is therefore less than $0.05|\Delta V/2|$.

C. Communication Power and Interference Power

Now we can define power in terms of these components: the interfering leaked power $P_{\rm leak}$ and the useful backscattered modulation $P_{\rm bs}$. Assume a detector with impedance Z is attached to the switched modulation. Time-averaged leaked power is

$$\text{Power delivered into } Z = \frac{1}{\mathrm{Re}(Z)} \frac{1}{T_2 - T_1} \int_{T_1}^{T_2} |v_{\rm leak}(t)|^2 dt \tag{11}$$

where the bounds $[T_1, T_2]$ represent the start and end of either the measurement period or the communication. We assume $T_2 - T_1 \gg 1/f_{\rm c}$, so the dc bias from truncated carrier cycles is neglected. Integrating the sinusoidal interference from (1)

$$P_{\rm leak} = \frac{|V_{\rm leak}|^2}{\mathrm{Re}(Z)}. \tag{12}$$

Similarly, the total carrier power P_{cw} is

$$P_{\rm cw} = \mathrm{PSD}[v(t)](f_{\rm c}) = \mathrm{PSD}[V(t)](0) = \frac{|\overline{V}|^2}{\mathrm{Re}(Z)}. \tag{13}$$

Importantly, $P_{\rm leak} = P_{\rm cw}$ only if $V_{\rm bs}(t)$ is defined as BPSK. ASK-defined $V_{\rm bs}(t)$ has its own carrier component, so the leaked power is different from $P_{\rm cw}$.

When the received baseband is defined as BPSK, signal magnitude in either state is $|\Delta V/2|$, so the same reasoning gives

$$\text{Received BPSK modulation power} = \frac{1}{4}\frac{|\Delta V|^2}{\mathrm{Re}(Z)} \tag{14}$$

which is the normalization asserted by [4] and [8].

The baseband receiver dc component in ASK modulation, $\overline{V}_{\rm bs} = \Delta V/2$, adds to the BPSK power to give the alternative normalization

$$\text{Received ASK modulation power} = \frac{1}{2}\frac{|\Delta V|^2}{\mathrm{Re}(Z)} \tag{15}$$

as in [6] and [7].

D. Energy Conservation at Network Interfaces

Energy conservation must apply in any meaningful physical model and can help estimate loose upper bounds for received power.

Assume we have a generator/detector with impedance Z attached to a arbitrary passive load Z_L. The incident and reflected power waves $\tilde{a}$ and $\tilde{b}$ at this interface can defined as [15]

$$\begin{aligned} \tilde{a} &= \frac{V}{2\sqrt{\mathrm{Re}(Z)}}\left(1 + \frac{Z}{Z_L}\right) = \frac{V_G}{2\sqrt{\mathrm{Re}(Z)}} \\ \tilde{b} &= \frac{V}{2\sqrt{\mathrm{Re}(Z)}}\left(1 - \frac{Z^*}{Z_L}\right) = \frac{V_G}{2\sqrt{\mathrm{Re}(Z)}}\frac{Z_L - Z^*}{Z_L + Z}. \end{aligned} \tag{16}$$

Voltage V is the same as in the previous section, and V_G is the Thévenin RMS source phasor. These normalizations give $P_{\rm av} = |\tilde{a}|^2$ as available power from the generator, $|\tilde{b}|^2$ as the reflected power absorbed back into Z, and $(P_L = |\tilde{a}|^2 - |\tilde{b}|^2)$ as power absorbed by Z_L. The expression for $\tilde{b}$ includes power-wave reflection coefficient $\tilde{\rho} = (Z_L - Z^*)/(Z_L + Z)$, which is bound by $|\tilde{\rho}| < 1$.

Let Z_L switch between Z_{L1} and Z_{L2}, with corresponding $\tilde{\rho}_1$ and $\tilde{\rho}_2$. The duty cycle from the previous section is 50%, so the time-averaged power absorbed by each load is $P_{L1}/P_{\rm av} = 0.5(1 - |\tilde{\rho}_1|^2)$ and $P_{L2}/P_{\rm av} = 0.5(1 - |\tilde{\rho}_2|^2)$. Inserting (13)–(15) into (16) gives reflected modulation and carrier power including matching effects

$$\frac{P_{\rm bs}}{P_{\rm av}} = \begin{cases} \frac{\frac{|\Delta V|^2}{(4{\rm Re}(Z))}}{\frac{|V_{\rm av}|^2}{{\rm Re}(Z)}} = \frac{1}{4}|\tilde{\rho}_1 - \tilde{\rho}_2|^2 & \text{(BPSK)} \\ \frac{\frac{|\Delta V|^2}{(2{\rm Re}(Z))}}{\frac{|V_{\rm av}|^2}{{\rm Re}(Z)}} = \frac{1}{2}|\tilde{\rho}_1 - \tilde{\rho}_2|^2 & \text{(ASK)} \end{cases}$$
$$\frac{P_{\rm leak}}{P_{\rm av}} = \begin{cases} \frac{|\overline{V}|^2}{{\rm Re}(Z)} = \frac{1}{4}|\tilde{\rho}_1 + \tilde{\rho}_2|^2 & \text{(BPSK)} \\ \frac{|V_1|^2}{{\rm Re}(Z)} = |\tilde{\rho}_1|^2 & \text{(ASK)}. \end{cases} \tag{17}$$

This is valid at any interface, including: 1) a tag chip-to-antenna bond (if parasitics are accounted for); 2) between a reader and its antenna (where propagation loss can make $|\tilde{\rho}_1 - \tilde{\rho}_1|^2$ quite small); or even 3) if a reader is attached directly to a switch.

Compare total power dissipated into Z and $Z_{L1,L2}$ against total available power for BPSK-defined $P_{\rm bs}$ and $P_{\rm leak}$

$$\begin{aligned}\frac{P_{\rm leak}}{P_{\rm av}} + \frac{P_{\rm bs}}{P_{\rm av}} + \frac{P_{L2}}{P_{\rm av}} + \frac{P_{L1}}{P_{\rm av}} &= \frac{1}{4}|\tilde{\rho}_1 - \tilde{\rho}_2|^2 + \frac{1}{4}|\tilde{\rho}_1 + \tilde{\rho}_2|^2 \\ &\quad - \frac{1}{2}(|\tilde{\rho}_1|^2 + |\tilde{\rho}_2|^2) + 1 \\ &= \frac{1}{2}|\tilde{\rho}_1\tilde{\rho}_2|[\cos(\angle\tilde{\rho}_1 - \angle\tilde{\rho}_2) \\ &\quad - \cos(\angle\tilde{\rho}_1 - \angle\tilde{\rho}_2)] + 1 \\ &= 1. \end{aligned} \tag{18}$$

Available input power is equal to the total BPSK dissipated power, satisfying basic physics for any passive $Z_{L1,L2}$ and Z.

At the switch generating the modulation, we can define "BPSK loads" as $|\rho_1| = |\rho_2|$ and "ASK loads" as $\rho_2 = 0$. This is independent of the terms "ASK" and "BPSK" applied to $P_{\rm bs}$ and $P_{\rm leak}$ at the detector. Passive EPC tag chip impedances are closer to "ASK loads": $\tilde{\rho}_2$ absorbs and harvests from $P_{\rm av}$ as a power supply, and a shorted $\tilde{\rho}_1$ reflects power. An EPC tag's Miller- or FM0-encoded ASK loads can deliver half of $P_{\rm av}$ into Z_{L2}[16]. From (17), the ASK loads result in received BPSK signal power of $P_{\rm bs} = P_{\rm leak} = P_{\rm av}/4$. A quarter of the incident energy is "wasted" here on the useless $P_{\rm cw}$.

For ASK-defined $P_{\rm bs}$ and $P_{\rm leak}$, dissipated and available energy are related by

$$\begin{aligned}&\frac{P_{\rm cw}}{P_{\rm av}} + \frac{P_{\rm bs}}{P_{\rm av}} + \frac{P_{L2}}{P_{\rm av}} + \frac{P_{L1}}{P_{\rm av}} \\ &\qquad = 1 + \frac{1}{2}|\tilde{\rho}_1 - \tilde{\rho}_2|^2 - \frac{1}{2}(|\tilde{\rho}_1|^2 - |\tilde{\rho}_2|^2). \end{aligned} \tag{19}$$

For the ideal "ASK loads" example, Z_{L2} still absorbs $P_{\rm av}/2$. The ASK signal power is also $P_{\rm av}/2$ (this includes the $P_{\rm av}/4$ carrier power that is separated as $P_{\rm leak}$ for BPSK signal power). For ideal "BPSK loads," however, there is no carrier power in the reflected signal, but the ASK modulation by definition assumes that there is. Thus, the ASK signal definition via (19) gives the $3P_{\rm av}$ reflected and absorbed by Z, which violates passivity.

As a final note, a third normalization of $P_{\rm bs}$ has also been borrowed from (32) in Green's 1963 thesis [17]. By incorporating carrier components into $P_{\rm bs}$, it also fails the passivity condition, giving $P_{\rm bs}/P_{\rm av} = 4$ for "BPSK loads."

IV. Conclusion

The authors recommend assuming that received digital backscatter is BPSK as (14), so the passivity condition (18) applies to any network plane between passive impedances.

A convenient side effect of the BPSK definition is that the interfering "leaked" carrier is the same as the carrier component measured in a received spectrum.

In contrast, ASK may cause nonphysical measurement and simulation results when testing battery-assisted backscattering transponders that use "BPSK loads."

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