

Reference Modulation for Calibrated Measurements of Tag Backscatter

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Abstract—This paper presents an approach for calibrating backscattering measurements from 860-960 MHz Ultra-High Frequency Radio Frequency Identification (UHF RFID) tags. An S-parameter model is formulated to relate diode switch and antenna input circuit parameters with the scattering performance of the calibration device. Measurements of modulated backscattered power agree with the model to within ± 0.1 dB. Tag backscatter measurements can then be calibrated by comparing them to the reference signal. In an example testbed, the expanded uncertainty of these measurements is estimated to be ± 0.4 dB, compared with uncertainties worse than -0.9 dB, $+1.2$ dB for methods that calibrate against radar cross section (RCS) standards in the same testbed.

I. INTRODUCTION

Continued technical advances in RFID systems are allowing passive transponders (“tags”) to operate with increasingly lower powers received from transceivers (“readers”) in UHF 860-960 MHz bands. This trend enables operation at greater range, but also means that backscattered replies from tags are becoming fainter. If this trend continues, existing interference effects [1] will be made worse, and reader detection of tags may become a dominant constraint on communication in practical deployments [2]. Unfortunately, performance test standards are immature for tags, and do not yet exist for readers. The lack of strong test standards may add effort and expense to the designing robust inventory tracking systems.

System operation, in contrast, has been more extensively standardized. Typical commercial systems comply with the EPC Global “UHF Class 1 Gen 2” and ISO 18000-6C standards, which outline the operation and parameters of the half-duplex protocol used for UHF RFID. In the forward link, a reader transmits a modulated carrier towards a tag field, which harvests the incident energy to supply power for communication and processing circuitry. Each tag responds by time-varying its antenna load impedance, encoding information in the backscattered carrier. This modulated backscattering implementation, known as load modulation, forms the return link. Similar processes have been used for antenna measurements [3] and espionage [4].

Existing test standards leave significant unresolved challenges, particularly in the return link. They do not yet address either reader sensitivity or interference rejection, though previous work has suggested that reader sensitivity is becoming a more significant system range constraint [2], and that readers are more vulnerable to interference than tags [1]. Standard ISO 18047-6 [5] prescribes a tag backscattering performance test characterized as the difference between the radar cross section values between its two load modulation states. The prescribed test method calibrates measurements of tag backscattering against the change in received power from adding a thin rod to the test environment.

The measurement uncertainty of one test performed this way was estimated to be approximately 2 dB [6]. The current ISO/IEC 18047-6 method introduces further systemic error by directly subtracting power quantities, neglecting phase effects, though existing works show how phase can be considered correctly [7][8][9]. Adding and removing the entire thin rod calibration standard introduces systemic error via structural-mode modulation, which interacts with multipath in the test environment differently [2] than tags’ antenna-mode modulation [7][8]. The use of such an electrically small calibration target requires faith in the accuracy of the analysis used to compute its RCS, which makes the measurement result not traceable to fundamental physical standards of any national metrology laboratory. These errors may make measurement results challenging to repeat between different testbeds, and as a result some parties may choose not to undertake the expense of running the tests.

To address these shortcomings, a new model and calibration device for traceable, more accurate measurements of backscattered tag power are proposed in this paper. The contributions are organized as follows:

- Section II introduces the model, based on S-parameters;
- Section III shows how a calibration device can be realized with a horn antenna and a laboratory switch, and characterized with the developed model parameters;
- Section IV presents the technique for calibrating tag backscatter with the calibration device;
- Section V shows a calibrated tag backscatter measurement with the new model and device.

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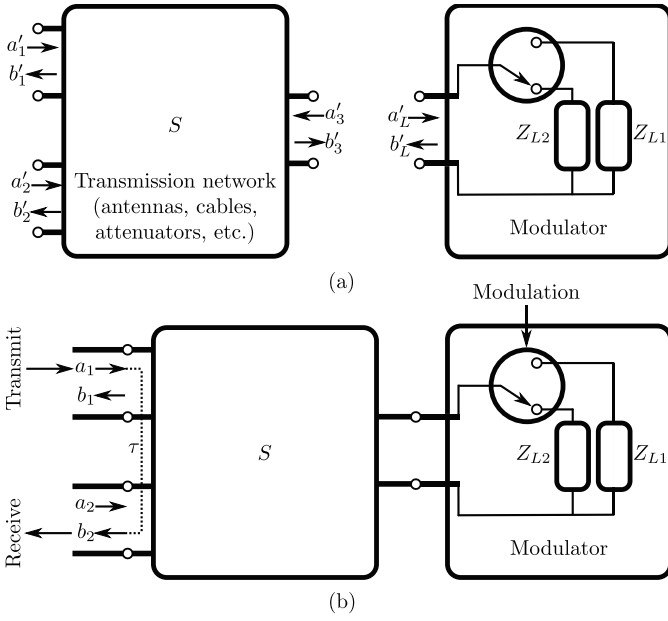


Fig. 1. Reference backscattering model illustrating S-parameters with the modulator (a) disconnected and (b) connected to the transmission network. When connected, a transmitter delivers a traveling wave a_1 into port 1 of an arbitrary transmission system S , which is described in (a). The reflection coefficient presented to waves arriving at port 3 switches between $\Gamma_L \rightarrow \Gamma_{L1}$ (the Z_{L1} state) and $\Gamma_L \rightarrow \Gamma_{L2}$ (the Z_{L2} state). Reflection back to port 2 is the traveling wave b_2 . The transmit-to-receive transmission coefficient in a tag state is $\tau = b_2/a_1$.

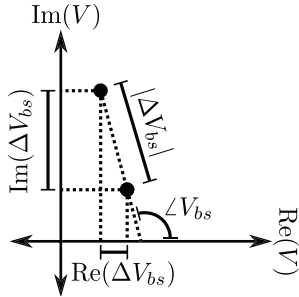


Fig. 2. Illustration of demodulated ΔV_{bs} on the complex (IQ) plane received from a reference backscatterer. The signal is shown in rectangular coordinates as $\text{Re}(\Delta V_{bs}) + j\text{Im}(\Delta V_{bs})$, and in polar form as magnitude and phase.

II. LINK AND LOAD MODULATION MODEL

A. Derivation

The proposed model for transmission and backscattering between an interrogator, and modulator (realized in subsequent sections as a calibration device) is illustrated in Fig. 1. The linear and reciprocal transmission network is shown as S , where interrogation signals are incident at port 1, backscattering is measured at port 2, and the backscatter modulator loads port 3. This network can incorporate any cables, antennas, linear propagation effects, or other linear circuit elements in the testbed.

If the two networks are disconnected (Fig. 1a), transmission coefficients between the loaded ports in the reciprocal trans-

mission are $S_{23} = S_{32} = b'_3/a'_2$ and $S_{31} = S_{13} = b'_3/a'_1$, and can be measured conveniently with a network analyzer if they are connectorized. Port 3 bisects the reference backscatter standard. The reflection coefficients of the backscattering antenna and the modulator are $S_{33} = b'_3/a'_3$ and $\Gamma_L = b'_L/a'_L$, respectively. These coefficients are measurable with a network analyzer by disconnecting the reference backscattering antenna from the modulator.

When the two networks are connected (as in Fig. 1b), an incident carrier wave from port 1 arrives at the load modulator, which is time-varying its impedance between two states. Reflections off of this interface arrive at port 2 with two corresponding states that have in-phase and quadrature (IQ) voltage components illustrated in Fig. 2.

The modulated signal ΔV_{bs} is related to a transmit signal V_{tx} in an arbitrary environment as follows. The traveling waves a_1 and b_2 in Fig. 1b are $a_1 = V_{tx}/Z_0^{1/2}$ and $b_2 = V_{bs}/Z_0^{1/2}$, with purely real port impedances Z_0 . An expression for the received wave b_2 in this system with port 3 loaded by the tag is [10]

$$b_2 = \left(S_{21} + \frac{S_{31}S_{23}\Gamma_L}{1 - S_{33}\Gamma_L} \right) a_1 + \left(S_{22} + \frac{S_{32}S_{23}\Gamma_L}{1 - S_{33}\Gamma_L} \right) a_2. \quad (1)$$

Assuming that Port 2 is not transmitting ($a_2 = 0$), the transmission coefficient $\tau = b_2/a_1$ between the transmit and receive ports becomes

$$\tau = S_{21} + \frac{S_{31}S_{23}\Gamma_L}{1 - S_{33}\Gamma_L}. \quad (2)$$

This relationship has been observed for other two-port modulated scattering models [11] [12]. If the modulator load alternates between $\{Z_{L1}, Z_{L2}\}$, corresponding to reflection coefficients $\Gamma_L \rightarrow \{\Gamma_{L1}, \Gamma_{L2}\}$, then the change in the transmission coefficient $\Delta\tau$ will be

$$\Delta\tau = \frac{\Delta V_{bs}}{V_{tx}} = S_{31}S_{23} \frac{\Gamma_{L2} - \Gamma_{L1}}{(1 - S_{33}\Gamma_{L2})(1 - S_{33}\Gamma_{L1})}. \quad (3)$$

The term on the right gives a convenient definition for a return modulation depth

$$M = \frac{\Gamma_{L2} - \Gamma_{L1}}{(1 - S_{33}\Gamma_{L2})(1 - S_{33}\Gamma_{L1})} = \frac{\Delta\tau}{S_{31}S_{23}}, \quad (4)$$

relating system behavior through the transmission network ($\Delta\tau$, S_{31} , and S_{23}) with measurable parameters inside the backscatter reference (Γ_{L1} , Γ_{L2} , and S_{33}).

The relationship in (3) expressed with received modulation power P_{bs} between the two switched states at the reader, transmission link losses $|S_{31}|^2$ and $|S_{23}|^2$, and the carrier transmit power P_{tx} is[†]

$$\frac{P_{bs}}{P_{tx}} = \frac{|\Delta V_{bs}|^2/2Z_0}{|V_{tx}|^2/2Z_0} = |\Delta\tau|^2 = |S_{31}|^2|S_{23}|^2|M|^2. \quad (5)$$

[†] Power quantities given in this model are *available* power. Finding power *delivered* into a receiver or from a source must additionally account for mismatch between the impedances on either side of the network interface.

If the input propagation environment (an antenna or transmission line) is matched to an instrument's characteristic impedance Z_0 ($S_{33} = 0$), then the modulation index in (5) expressed in terms of power $|M|^2$ is bounded by $|M|^2 < 4$. This is the case of two modulator states with reflection coefficients of with magnitude 1 that are out of phase. If $S_{33} \neq 0$, $|M|$ can become quite large or small, but energy is conserved because the mismatch reduces the magnitude of the transmission coefficients S_{31} and S_{23} .

This relationship applies most directly to bistatic measurement antenna configurations, with separate transmit and receive antennas corresponding to ports 1 and 2 of Fig. 1. It can also apply to monostatic configurations if S incorporates a circulator.

B. Comparison with other scattering models

The best-known modulated tag scattering characteristic is known as either a “differential” (e.g. [7] [13]) or “modulated” ([14]) radar cross section (RCS). These terms are also sometimes used interchangeably with the terms “delta RCS” or “ Δ RCS” used in ISO 18047-6, though this term may cause confusion over the role of phase in tag scattering.

The modulated RCS model is formulated with power waves [7][13][14][15], which should not be confused with the S-parameter traveling waves used in this paper. The power wave parameters are convenient for chip-to-antenna matching in tag design. In metrology, however, power waves are not suitable for extracting physical parameters like impedance from measured data [16], making them unsuitable for calibration purposes.

A more recent model of tag scattering based on reciprocity and impedance (Z-) parameters was introduced in [8], and can be shown to be equivalent to the work presented here. Like ours, this model requires knowledge of a chip and antenna circuit parameters. The choice of S-parameters instead Z-parameters was motivated by practical because far-field propagation effects between antennas are understood in terms of traveling waves (rather than mutual impedances), and because microwave instrumentation uncertainty is expressed in terms of S-parameters.

This paper is not intended to offer a “black box” characterization like RCS for tag testing. Direct application of the model presented here to tag characterization would require chip and antenna parameters either by prior knowledge or destructive disassembly of a tag.

C. An idealized free-space example

Though the goal of this work is not to present RCS measurements, the relationship between the S-parameter scattering model and modulated RCS can give some insight. As a simple test case, consider an idealized monostatic system in an anechoic environment. The backscattering and detection antennas are well-matched and co-polarized, and we assume there are no near-field effects. Measurement and backscattering antennas have gains G and G_{bs} towards each other. With

antenna reciprocity ($|S_{31}|^2 = |S_{23}|^2$ in (5)) and the Friis transmission equation, the backscattered power is

$$\frac{P_{bs}}{P_{tx}} = G^2 G_{bs}^2 \left(\frac{\lambda}{4\pi R} \right)^4 |M|^2, \quad (6)$$

where R is the range between the measurement and backscattering antennas and λ is the free-space carrier wavelength. This makes the same assumptions as the radar equation, and has the same form, so it can be related to the modulated radar cross section σ_{mod} (see for example [14]) used for tags:

$$\sigma_{mod} = G_{bs}^2 |M|^2 \frac{\lambda^2}{4\pi}. \quad (7)$$

Relating the S-parameter network model to modulated RCS in free space therefore requires additional information about the gain of the backscattering antenna.

III. A REFERENCE SOURCE OF BACKSCATTER MODULATION

The model in the previous section suggests that a carefully characterized source of reference modulation with a known modulation depth ($|M|$) and measurements of link losses ($|S_{31}|$ and $|S_{23}|$) are sufficient to predict the backscattered reference power. This section introduces a device that can be characterized to serve this purpose, and how it can be validated as an accurate reference for calibrating tag backscatter.

A. Calibrated signal generation

The modulator shown in Fig. 3 realizes the model in Fig. 1. The switch has a nominal 20 ns rise/fall time to within 10% of steady state, which is fast enough to emulate the maximum 640 kHz symbol rate by tags compliant with ISO/IEC 18000-6C.

A standard gain horn with a measured gain of 6.69 dBi at 1 GHz is the reference backscattering antenna. It has return loss greater than 10 dB across 860-960 MHz, corresponding to $|S_{33}|$.

Instruments with 50 Ω input impedance serve both as a matched load modulation state (with $|\Gamma_{L2}| \approx 0$), and to allow measurements of interrogation signal link losses. A network analyzer makes a convenient matched load for characterizing the device, and a power meter is subsequently used to measure received interrogation power. An additional 3 dB pad between the switch output and the matched load attenuates reflections between the horn and the instrument. The other switch load is a short, for $|\Gamma_{L1}| \approx 1$, though the actual $|\Gamma_{L1}|$ is approximately 2 dB smaller because of switch insertion loss.

With $|\Gamma_{L2} - \Gamma_{L1}| \approx 1$, and the backscatter antenna approximately matched for small $|S_{33}|$, the anticipated $|M|$ is near 0 dB.

B. Measuring modulation depth

To become suitable for calibrations with reference backscatter, the modulation depth must be measured accurately. Equation (4) gives a choice between two sets of parameters that can be measured with a network analyzer: “wireless”

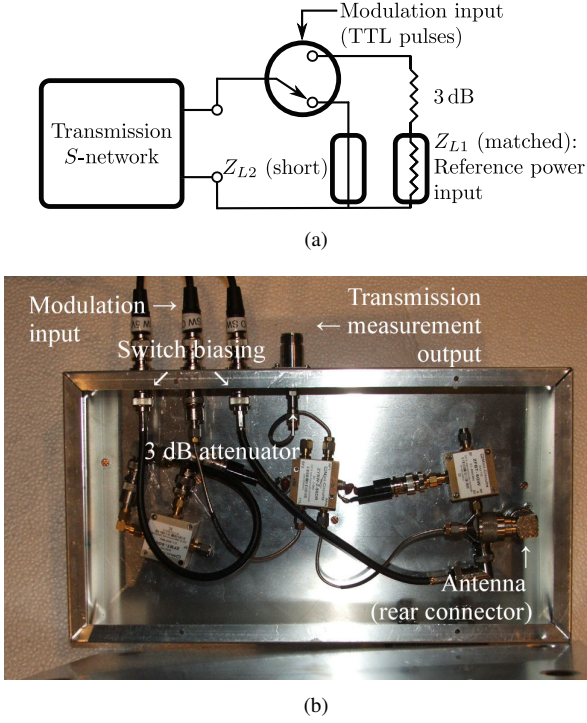


Fig. 3. Simplified schematic (a) and realization (b) of a modulation switch for reference backscatter calibrations. The load Z_{L1} is intended to connect with a matched 50Ω instrument such as a power sensor or network analyzer, to measure power delivered to the backscatter reference and serve as a matched reflection state for modulation. The device is mounted in a $33\text{ cm} \times 18\text{ cm} \times 5\text{ cm}$ shielded box with $\pm 5\text{ V}$ DC biasing inputs, and bias tees to improve DC to RF isolation.

measurements of S_{23} , S_{31} , and $\Delta\tau$, or reflection coefficient measurements of Γ_{L2} , Γ_{L1} , and S_{33} . In either case, the modulation state of the modulator is fixed with a DC voltage supply during measurement.

One advantage of calibrating M from measurements of Γ_{L2} , Γ_{L1} , and S_{33} directly is that (to first order) measurement dynamic range is not reduced by moving the reference backscatter antenna in an anechoic test environment. Further, Γ_{L2} and Γ_{L1} can be measured with phase-stable cables near the network analyzer more accurately than with the long cables that are necessary to measure reflection coefficients of objects inside the test chamber.

Calibrating M from $\Delta\tau$ and measurements of propagation losses S_{23} and S_{31} has different advantages. Transmission measurements of propagation losses can have smaller uncertainties than the reflection measurements in the first approach, but any motion in the long cables may introduce additional phase errors. This calibration also needs fewer measurements, which may reduce the contribution of operator mistakes to measurement error.

Detailed quantitative comparison of uncertainties in these approaches will be left for future work. The following subsection will validate that either set of measurements can produce a valid characterization of $|M|$.

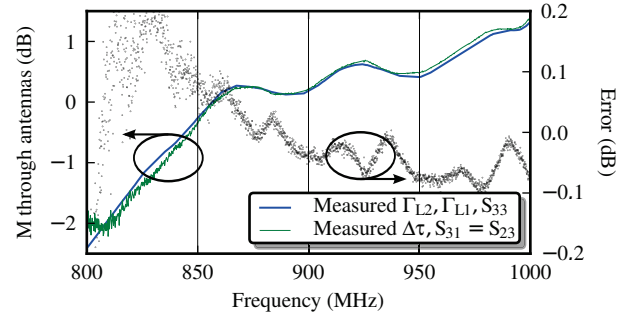


Fig. 4. Validation of the reference backscatter with a network analyzer in a semi-anechoic test environment, computed with measurements of the terms in (4). The curves agree to $\pm 0.1\text{ dB}$ over the 860-960 MHz tag response bandwidth.

C. Validation

Because the model in equation (4) gives two expressions for M in terms measurable network parameters, the modulation depth characterization can be validated by measuring the parameters for each with a network analyzer and comparing the results.

The validation tests, for simplicity, were performed in a mono-static. The transmit and receive antenna is a commercial RFID patch with at least 10 dB return loss across 895-940 MHz. The change in reflection coefficient $\Delta\rho$ into the antenna was taken to be equivalent to $\Delta\tau$, with the simplification $S_{23} = S_{13}$. The anechoic environment reduces interference from outside signals, but the calibration for backscattered signal levels applies in other, more reflective environments too, if interference is below a tolerable level.

Results are shown in Fig. 4. Across the 860-960 MHz tag response bandwidth, the two measurements of M agree within 0.1 dB. Below 860 MHz, detection antenna mismatch introduces additional noise in transmission measurements S_{31} and $\Delta\rho$, because received signals are weaker.

IV. CALIBRATION WITH REFERENCE MODULATION

A. Calibration procedure

1) *Reference $|M|$ calibration:* The validated characterization of the reference backscatter modulation depth M from the previous section is used. In this work, tag calibrations use the modulation depth based on measurements of Γ_{L2} , Γ_{L1} , and S_{33} .

2) *Transmission loss measurements:* During tag measurements, it is impractical to measure transmission coefficients $|S_{31}|^2 = |S_{13}|^2$ and $|S_{23}|^2 = |S_{32}|^2$ with a network analyzer. Instead, we use power sensors to measure (1) transmitted interrogation power P_{tx} available to port 1 or 2 with a coupler, and (2) power received at the output of the modulator switch, $P_{rx}^{(ref)}$. In this paper, transmit, received (by transmission), and backscattered power from an interrogation into port n are represented as $P_{tx,n}$, $P_{rx,n}$, and $P_{bs,n}$. Assuming the network analyzer and power sensors are similarly well-matched, power

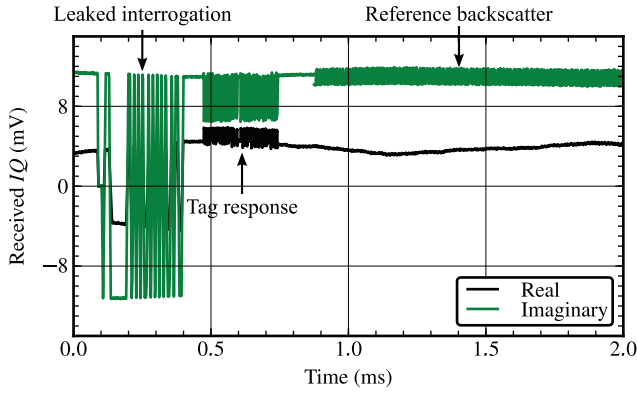


Fig. 5. A demodulated trace from a transaction at 910 MHz with an ISO/IEC 18000-6C tag received by a spectrum analyzer. It shows leaked interrogation modulation from the forward link, the tag response from the reverse link, and reference backscatter from the calibration device introduced in this paper. In use, the reference backscatter is only turned on when it is being measured, to avoid interfering with the tag.

measurements and transmission losses are related with

$$|S_{31}|^2 = |S_{13}|^2 = \frac{P_{rx,1}^{(ref)}}{P_{tx,1}} \quad (8)$$

and

$$|S_{32}|^2 = |S_{23}|^2 = \frac{P_{rx,2}^{(ref)}}{P_{tx,2}}. \quad (9)$$

$P_{rx}/P_{tx,1}$ and $P_{rx}/P_{tx,2}$ are taken as $|S_{32}|^2$ and $|S_{31}|^2$. Both sensors are configured to measure average power during the period after the tag reply while the interrogation power is left on. This period was set to 1 ms, which is longer than that of typically deployed readers to reduce noise by averaging.

Loss in the switch reduces the measured power compared to the available $P_{rx}^{(ref)}$ out of the backscattering antenna. To “back out” $P_{rx}^{(ref)}$, the full two-port scattering parameters of the switch are used to de-embed the power available out of port 3 with transfer (T-) parameters.

3) *Backscattered signal measurements and processing:* Backscattered power $P_{bs}^{(tag, meas)}$ and $P_{bs}^{(ref, meas)}$ are measured with IQ demodulation on a spectrum analyzer. Example signal traces are shown in Fig. 5. The interrogation signal is an ISO/IEC 18000-6C query command, with tag and reference responses modulated at 160 kHz.

Tag responses to query requests are measured in a 240 μ s period, and reference backscatter in 1 ms. Connecting the 10 MHz frequency reference from the spectrum analyzer to the RFID interrogation source, carrier phase drifted by less than 1.3° per symbol at 160 kHz modulation, introducing negligible error to measurements of signal levels between adjacent symbols. Across entire tag or reference signal traces, however, the carrier drifts by up to 13° . This made separating the signal into the two states in software (“clustering”) more challenging: slow drift in each of the I and Q components was often larger than the backscattered signal, making the straightforward histogram analysis suggested by IEEE 181 [17] unsuitable.

Instead, states are clustered by applying a Gaussian filter to the I and Q scattering components, and finding sharp peaks after differentiating the result. These peaks are reported when they are a local amplitude maximum; this threshold was set somewhat arbitrarily to the standard deviation of the filtered signal. The Gaussian filter is convenient here because it removes noise without shifting pulse edges; it is a simpler one-dimensional implementation of Canny edge detection [18].

If ΔV_I and ΔV_Q are the differences between the averaged two backscattering state measurements of each component, ΔV_{bs} is computed as (3),

$$\Delta V_{bs} = (\Delta V_I + j\Delta V_Q) \exp j\phi, \quad (10)$$

where the exponential term accounts for the arbitrariness I - Q plane rotation. For vector voltage quantity ΔV_{bs} , averaging across neighboring state transitions in different frames reduces measurement noise, potentially to below -100 dBm.

The power in these two states, corresponding to P_{bs} from (5), was calculated as

$$P_{bs} = \frac{|\Delta V_{bs}|^2}{2Z_0} = \frac{|\Delta V_I + j\Delta V_Q|^2}{2Z_0}, \quad (11)$$

in agreement with previous work [7][9][13][19].

4) *Correcting the tag backscatter measurement:* Assuming bandwidth of all backscattered signals are narrow about the interrogating carrier, and that cable and antenna mismatch and losses are linear with power, the fractional power lost will be the same for both a tag and reference backscatter:

$$\frac{P_{bs}^{(ref, meas)}}{P_{bs}^{(ref)}} = \frac{P_{bs}^{(tag, meas)}}{P_{bs}^{(tag)}}. \quad (12)$$

This can be rearranged to find “true” backscattered power received from the tag,

$$P_{bs}^{(tag)} = \frac{P_{bs}^{(tag, meas)}}{P_{bs}^{(ref, meas)}} P_{bs}^{(ref)}. \quad (13)$$

Equations (5), (8), and (9) can be substituted for $P_{bs}^{(ref)}$, so for interrogation through port 1,

$$P_{bs,1}^{(tag)} = \frac{P_{bs,1}^{(tag, meas)}}{P_{bs,1}^{(ref, meas)}} \frac{P_{rx,2}^{(ref)}}{P_{tx,2}} P_{rx,1}^{(ref)} |M|^2. \quad (14)$$

or through port 2,

$$P_{bs,2}^{(tag)} = \frac{P_{bs,2}^{(tag, meas)}}{P_{bs,2}^{(ref, meas)}} \frac{P_{rx,1}^{(ref)}}{P_{tx,1}} P_{rx,2}^{(ref)} |M|^2. \quad (15)$$

Notice that, like the RCS calibration in ISO 18047-6, effects of mismatch or cable losses between the receive antenna and the measurement instrument are removed in the calibration process.

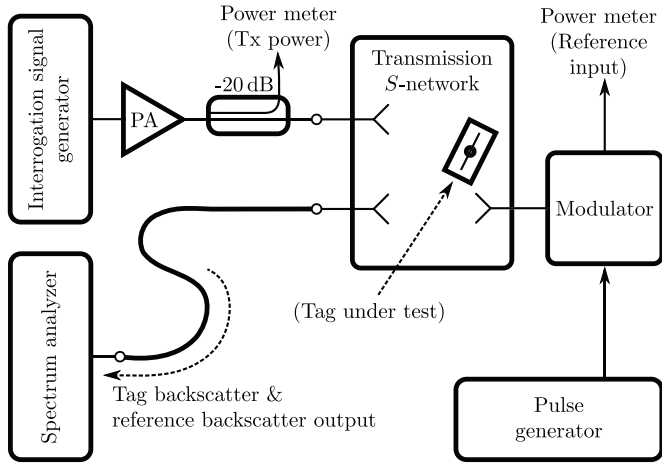


Fig. 6. Setup for calibrated tag backscattering measurements. Measurements were automated by computer via general purpose interface bus (GPIB).

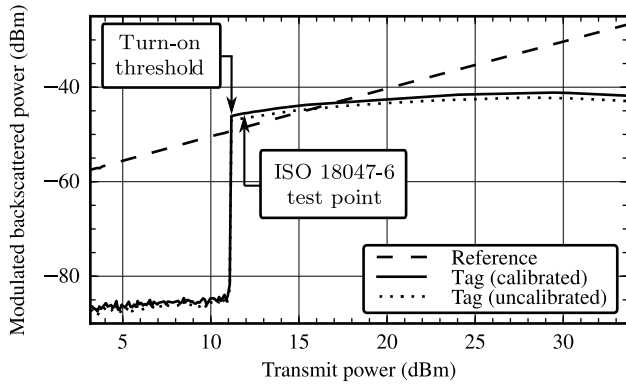


Fig. 7. Measurements of backscattered power comparing raw tag and reference signals with the calibrated measurement of the tag. The reference was switched at 160 kHz modulation, and the tag was interrogated with signal parameters described in table I. Signals from each were measured with IQ demodulation as in Fig. 8. The calibration used M calculated with equation (4) from measurements of Γ_{L1} , Γ_{L2} , and S_{33} .

V. EXAMPLE CALIBRATIONS

A. Calibrated tag measurements

With the procedure established, we next present an example tag backscatter measurement, calibrated with the backscattering source. The overall test setup is summarized in Fig. 6 and Table I. Instruments are controlled with computer via general purpose interface bus (GPIB).

The parameters from Table I were used in the calibrated tag backscatter measurement shown in Fig. 7, showing highly nonlinear raw and calibrated measurements of the tag under test with the reference backscatter for comparison. Reference backscatter was turned off during measurements of the tag response, to prevent interference.

B. Measurement errors and uncertainty

To help gauge the effectiveness of the reference backscatter calibration, an uncertainty estimate is presented in Table II,

TABLE I
ISO/IEC 18000-6C TAG QUERY PARAMETERS

Reader-to-tag modulation	PR-ASK
Tag-to-reader modulation	FM0
Tag-to-reader link rate (BLF)	160 kHz
Reader-to-tag link rate	160 kHz (data 0) 91 kHz (data 1)
Anticollision slots (Q)	0 (no slots)
Delay after tag response [†] (T2)	1 ms
Tari	6.25 μ s

[†] Reference backscattering was measured during this period

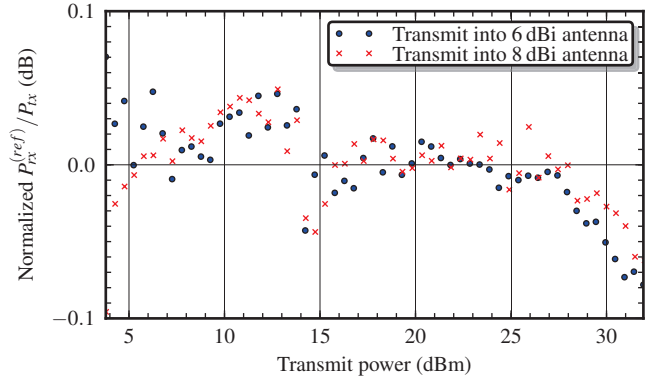


Fig. 8. Testbed linearity tests, performed by sweeping the power of the interrogating wave and measuring the reference backscattered power. The backscatter reference load-modulated 910 MHz carrier reflections at 160 kHz with the circuit described in Fig. 3. Deviation from linearity below 32 dBm input power was less than 0.1 dB.

with a comparison against other measurement approaches taken in the same testbed. Table IIb, for example, estimates uncertainty in hypothetical backscattered power measurement based on the spectrum analyzer's built-in alignment, with cable losses measured and de-embedded. The measurement in Table IIc is meant to reflect the uncertainty of a backscattering measurement that has been carefully calibrated with a thin $\lambda/2$ rod as a scattering standard, based in part on results from [2]. All of these measurements are assumed to use IQ demodulation on the same spectrum analyzer.

Testbed linearity is estimated from measurements of the reference backscatter as a function of power, shown in Fig. 8. This measurement was performed with a bistatic receiver setup. Results between the two traces agreed within ± 0.1 dB, which is then the estimate for linearity and noise uncertainty.

Reported values follow the methods for evaluating uncertainty expressed in [20] as closely as possible. Each error source is assumed to have a zero mean normal distribution, except multiple reflection errors that are assumed to have a U-shaped distribution. Supplied error estimates are type B, except the noise and linearity term, estimated from 8, and the IQ level measurement errors, estimated by simulation of noisy bandlimited pulses.

These uncertainty terms, because they are shown in decibels, are expressed relative to the actual measurement value. Each is listed as an expanded uncertainty, and corresponds to a fractional term u_i that is the contribution of the individual

TABLE II
EXPANDED UNCERTAINTY ESTIMATES (COVERAGE FACTOR 2)

(a) <i>Backscattered tag signal measurement (with reference backscatter calibration)</i>	
<i>M</i> calibration	±0.25 dB
Power measurements	±0.25 dB
Multiple reflections	±0.15 dB
<i>IQ</i> level measurements	±0.05 dB
Noise and nonlinearity	±0.1 dB
Expanded uncertainty	±0.4 dB
(b) <i>Backscattered tag signal measurement (without scattering calibration)</i>	
Transmit power measurements	±0.15 dB
Measurement cable losses	±0.1 dB
Instrument uncertainty	±0.9 dB
<i>IQ</i> level processing	±0.15 dB
Noise and nonlinearity	±0.1 dB
Expanded uncertainty	−1.2 dB, +0.9 dB
(c) <i>Backscattered tag signal measurement (with ISO 18047-6 λ/2 rod calibration)</i>	
Reference λ/2 rod RCS	(unknown)
Multiple reflections	±0.9 dB [2]
<i>IQ</i> carrier phase drift	±0.23 dB
<i>IQ</i> level processing	±0.15 dB
Noise and nonlinearity	±0.1 dB
Expanded uncertainty	−1.2 dB, +0.9 dB
(neglecting rod RCS uncertainty)	

error source as a fraction of the measured value. The reported value in decibels is related to u_i with

$$u_i \text{ (in dB)} = 10 \log_{10} (1 \pm u_i). \quad (16)$$

The expanded uncertainty estimate u_c from n terms is computed with

$$u_c \text{ (in dB)} = 10 \log_{10} \left(1 \pm \sqrt{\sum_{i=1}^n u_i^2} \right) \quad (17)$$

by the law of propagation of uncertainty [20], which applies here under the assumption that errors are zero-mean and that dominant errors are normally distributed. The sensitivity coefficients are already included in the itemized uncertainties listed in Table II.

Some errors are common to all of the measurements. The noise and nonlinearity uncertainty is estimated conservatively from (8). Uncertainty in the *IQ* measurement processing, caused by ringing or level clustering errors, is quoted as ±0.05 dB for results based on *relative* backscattering signal results (like the method presented in this paper), or ±0.15 dB for *absolute* measurements, based on simulation of noisy, band-limited pulses. Spectrum analyzer measurement uncertainty is contributes to errors in all of these cases, but when a scattering standard is used, the relative measurement result from the instrument is so small that it is negligible. This contrasts with the result in Table IIb, which is a dominant ±0.9 dB source of uncertainty estimated from the manufacturer's specification. Because the modulation of the λ/2 rod

standard in the measurement of Table IIc is performed by (very slowly) moving the standard by hand, the ±13° phase error becomes significant, and corresponds to a worst-case ±0.23 dB uncertainty.

The measurement technique presented in this paper has some unique uncertainty terms. Errors introduced by calibrations of *M* are a caused by reflection coefficient measurement uncertainty of the network analyzer and the short-open-load calibration standards. The corresponding error estimate is based on measuring verification load standards on the network analyzer after calibration, and on analysis of manufacturer uncertainty specifications. The power measurement uncertainty ±0.25 dB is also based on analysis of the manufacturer's power sensor specifications, from analysis of correlated and uncorrelated error terms between the two power measurements of (14) and (15). The multiple reflection estimate is a conservative estimate of detuning effects on the reference backscatter antenna reflection coefficient, caused by variations in the test setup, such as variations between scattering off of different tags under test. Empirical tests showed that reorienting the antenna or moving the tag in the test zone caused errors smaller than 0.1 dB.

VI. CONCLUSION

The reference backscatter modulation approach demonstrated here improves signal accuracy compared to radar cross section methods, with common lab equipment available at low cost from many different vendors.

With additional effort in analysis of the uncertainty of this technique as part of RCS tag tests, the authors hope it will become useful for future versions of tag performance test standards. Further work may also prove the effectiveness of the technique in small anechoic environments to reduce test costs. Reference signals may also become practical as in sensitivity or interference rejection tests.

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