A Microwave Noncontact Identification Transponder Using Subharmonic Interrogation

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***Abstract*—This paper presents the design and analysis of a novel microwave radio (RF/ID) transponder that has been developed for remote identification of personnel and articles, such as in "wireless key" entry systems. Based on a subharmonically pumped quasi-optical mixer, the transponder is activated by a C-band interrogation beam to upconvert and radiate a digitally modulated identification tone at two X-band frequencies. These response signals are nonharmonically related to the interrogation signal. The frequency conversion process is analyzed using a Volterra series model that includes the effects of self-bias current, coupled with a scattering metric that is directly useful in radio link calculations.**

I. Introduction

The need for automatic identification of articles and personnel has grown rapidly in recent years with the increased use of computerized systems for security and control tasks. The primary limitation of traditional schemes such as magnetically encoded cards, namely the requirement of physical contact between the card and reader, has provided impetus for the development of noncontact schemes where identification can be made at a distance. Noncontact identification schemes using radio frequencies (RF/ID systems) have several advantages over comparable optical systems, such as better penetration of obstructing materials (e.g., clothing, soot) and easier electronic manipulation of the identifying signals. Microwave frequencies in particular are attractive due to relatively low radio noise and interference levels, wide available bandwidth for high-speed data transfer, and physically small efficient antennas [1]. In a typical system, transponders (ranging from electronic ID badges to anti-theft tags) are read, or interrogated, by a microwave beam that causes them to emit a coded response. The response signal contains information that allows the interrogator to detect and identify the transponder in view. Various types of response signals are in use, including simple back-scatter with modulation in amplitude [2], phase [3] or both (i.e., SSB) [4], or with a controlled time delay [5]. In these systems the transponder responds on the same frequency as the interrogation signal, so the modulation imposed on the response serves both to transmit information from the transponder and to distinguish it from background clutter. Other transponders generate a harmonic [6], [7] or subharmonic [8] of the interrogation

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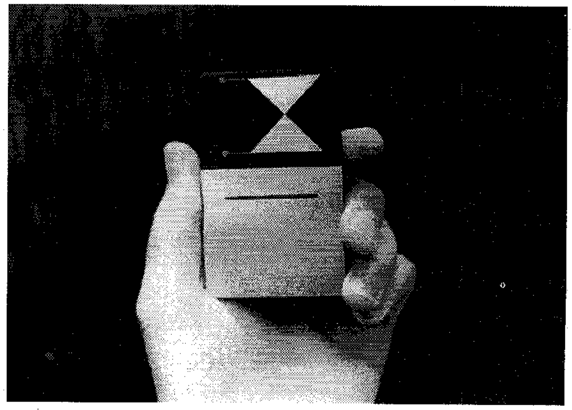

Fig. 1. The microwave noncontact identification card.

signal; this can be accomplished inexpensively using a diode or other nonlinear device and provides a response signal that is unique among the incidental backscatter from the linear objects in view.

In this paper we present a transponder in the form of an ID card (Fig. 1) that is interrogated at frequency ω_i, from which it upconverts a locally generated data signal at ω_d to the response frequencies $2\omega_i \pm \omega_d$ which are radiated: A 6-GHz interrogation beam illuminating the card will produce scattered responses at 11.990 and 12.010 GHz when the transponder's data carrier is preset to 10 MHz. The transponder is compatible with interrogation frequencies in the range 4–7 GHz, which includes the 5.8-GHz ISM band. The data carrier ω_d, hence the response signal, is ASK modulated with the identification code for the card—on the prototype, an 8-bit code, which is easily expandable to over 32 bits, was used. The advantages of this scheme include the creation of new microwave frequencies an octave apart from the interrogation signal, away from interference in the interrogation band, without the need for a microwave oscillator on the card. Unlike in harmonic-generation transponder systems, these frequencies are not harmonically related to the interrogation signal; therefore the problem of false detection resulting from an interrogator receiving reflections of its own transmitted harmonics is avoided. The carrier tone ω_d and the data sequence can be used independently or jointly to identify a card, where for instance a card group is identified by virtue of a common data carrier frequency. In an access control system, simple card readers could be used where only the card group, hence ω_d, need be determined such as at a building entrance, supplemented by readers located near internal doors that can detect the ASK code and obtain the detailed identity of the

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card. This scheme allows cards from different groups to be interrogated simultaneously without mutual radio interference. Further, with the addition of simple circuitry allowing the data carrier frequency to hop through a random sequence of n tones, numerous transponders could transmit their ID codes simultaneously in a dynamic frequency-division multiplex scheme, to be received by a multichannel or IFM-type receiver; with only four adequately spaced tones, such a system could accommodate several thousand transponders taken several at a time. Subharmonic interrogation is advantageous in that the benefits of operating at the higher frequency are accrued at the receiver (higher antenna gain, lower antenna noise, wider bandwidth) while requiring a high power source only at the lower frequency, which also makes the approach attractive for future millimeter-wave ID applications.

II. The Transponder Circuit

The transponder is based upon a quasi-optical structure first proposed by Stephan and Itoh for imaging arrays [9] and telemetry applications [10]. This structure consists of an antiparallel Schottky diode pair mounted at the terminals of a planar bowtie (triangular dipole) antenna (Fig. 2). The broadband nature of this structure allows it to receive the interrogation signal and transmit the response an octave apart. Since the ideal antiparallel diode pair has only odd terms in its nonlinear i-v curve, it generates only odd order mixing products in response to a multitone input; aside from the linear scattering of the interrogation signal, the dominant radiated product is the desired third-order term at $2\omega_i \pm \omega_d$. All even order currents, including the second harmonic $2\omega_i$ and dc, circulate only in the diode loop.

Generation of the transponder's data signal requires dc power, which must either be provided by an energy source on the card or extracted from the interrogation beam. In either case, detection of the interrogation signal is required to activate the card from a standby state. Ideally, the rectified dc current circulating in the mixer diode loop could be blocked with series capacitors and coupled out to power the transponder; both diodes would then be in series to provide improved dc output voltage compared to a single detector diode. However, for the low-frequency data signals used in the prototype, the required dc coupling network was judged to be impractical. Rather, a slot antenna is cut into the ground plane of the card, coupled to a microstrip line on the other side. The slot dimensions (length $1\lambda_g$, width 0.025λ at 6 GHz) were chosen to obtain a broadband low impedance, occurring around the second resonance, at the plane of the slot. A $\lambda/4$ microstrip line impedance-matches the slot to a Schottky diode to form a C-band receiver with a wide, bidirectional beam comparable to that of the bowtie. The detected dc signal activates the transponder from a power-conserving standby state. Low-power CMOS (74HC) was used in conjunction with discrete transistors on the prototype to generate a modulated, crystal-controlled 9.83 MHz data signal; standby and operating currents were 25 μA and 1 mA, respectively, from a +1.5 V silver oxide battery. Reducing the standby current to 1 μA could increase the 1–2 year battery life to 10 years, when interrogated 1 min. daily.

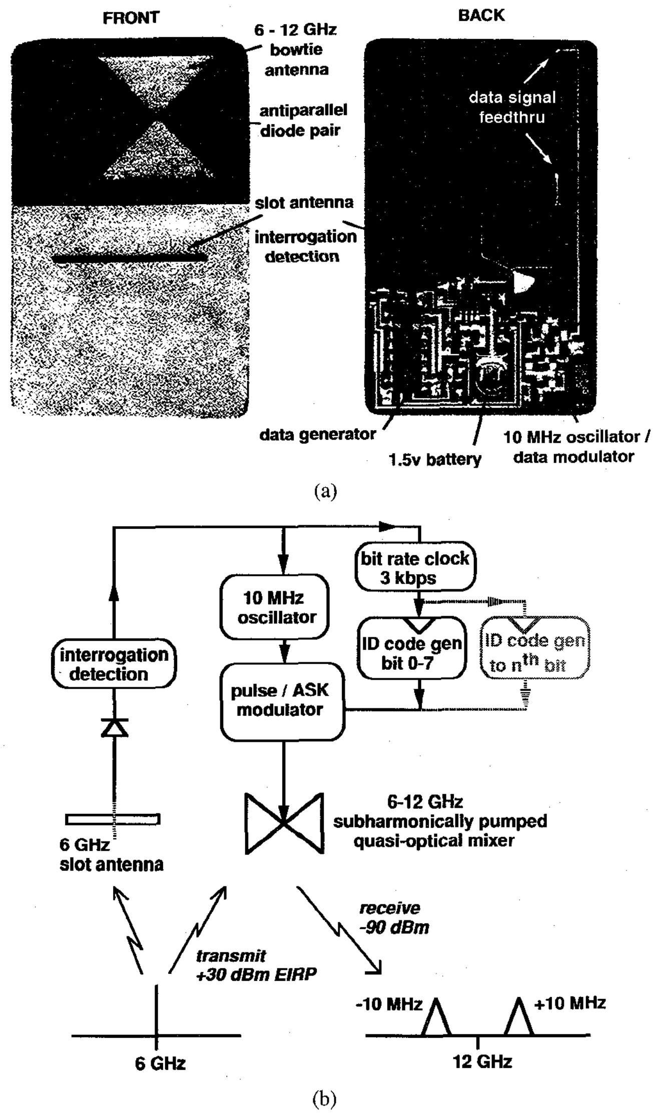


Fig. 2. (a) Elements of the transponder card. (b) Operational block diagram, showing the received and transmitted signals.

This could be effected by integrating the transponder circuitry into a single ASIC, which would allow far more sophisticated and programmable ID coding, and would open the possibility of obtaining power from the interrogation beam. In Fig. 3 the necessary interrogation EIRP is shown for a hypothetical transponder with various dc power requirements, given a 6-GHz slot/dipole receiver with $\eta = 20\%$ RF-DC conversion efficiency. For transponders requiring low dc power, on the order of microwatts, beam powering is clearly practical even at significant range (>3 m). As the power extracted from the beam is the product of incident beam intensity and effective ($\times\eta$) receiving aperture area, the dc power requirement translates to a minimum beam intensity that must be present during interrogation. If this minimum intensity exceeds the accepted limits of safe exposure to RF, the beam-powered transponder cannot be safely used for personnel ID at any range. Superior range is always obtained using a power conserving battery-powered card, minimizing the need for high-power RF sources.

In Fig. 4, the transponder is under interrogation from a test system at a distance of 5 ft. (1.5 m); the transmitted

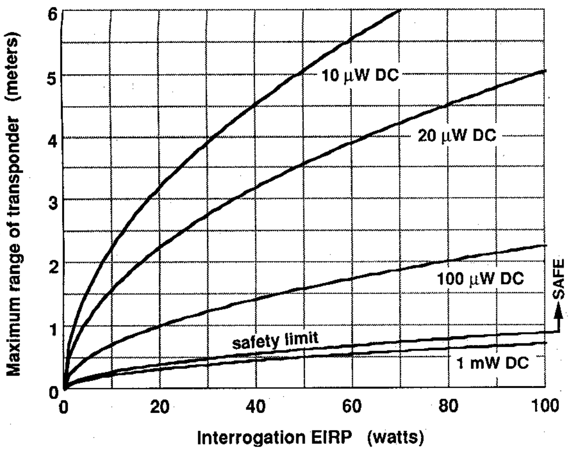


Fig. 3. Maximum usable range of transponder versus interrogation EIRP, for several dc power requirements. Hypothetical beam-powered transponder has a 6-GHz dipole receive antenna, 20% RF-DC conversion efficiency. Safety limit is US-ANSI 1 mW/cm^2.

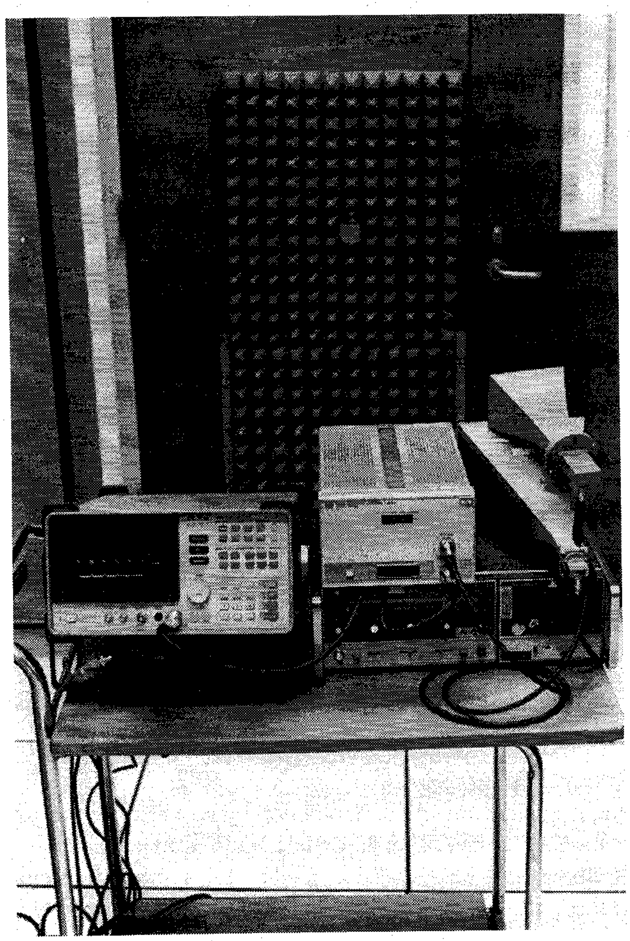
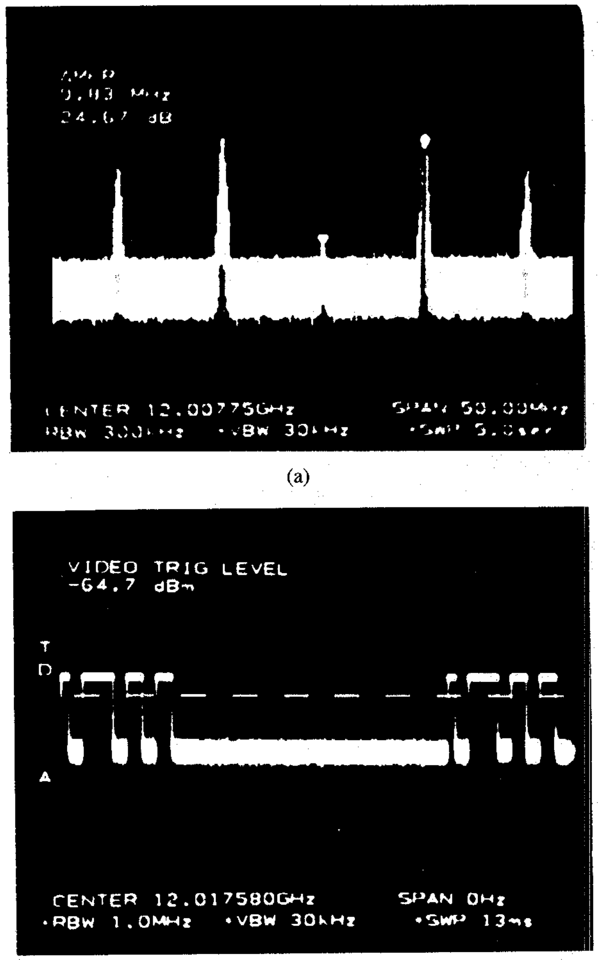


Fig. 4. Experimental interrogation test system.

power is 200 mW into a +15.5 dBi pyramidal horn (+38 dBm EIRP). An HP 8562A spectrum analyzer is used as a receiver, operating in zero span mode at the response frequency with a 1-MHz bandwidth. Fed by a +22 dBi pyramidal horn connected to a preamplifier, this system has a sensitivity of -105 dBm. Fig. 5 shows the scattered response signal, exhibiting adequate signal/noise ratio for noncoherent demodulation of the 8-bit ID code.



(a)

(b)

Fig. 5. (a) Spectrum and (b) demodulated upper sideband of response signal for a 6.003875-GHz interrogation; transponder's ID code is "10110101."

III. Quasi-Optical Mixer Design and Analysis

A. Mixer Design

The quasi-optical mixer comprises two diodes connected in antiparallel at the terminals of a bowtie antenna, illustrated in Fig. 6. An equivalent nonlinear circuit, Fig. 7(a), was analyzed using the harmonic balance technique to determine the effect of the embedding impedances presented to the diodes by the bowtie antenna and the data signal generator. A 90° flare angle was chosen for the bowtie, to approximate a self-complermentary antenna and minimize the variation of the feedpoint impedance Z_a with frequency. The experimental data of Brown and Woodward [11] was used to estimate this impedance, and the effect of the card substrate material (31 mil RT/Duroid 5870, $\varepsilon_r = 2.33$) was neglected. A Hewlett-Packard HCSH-5530 matched Schottky diode pair in a beam-lead T package was used (measured $I_s = 11.8$ nA, $\eta = 1.08, R_s = 5\,\Omega, C_{j0} = 0.1$ pF, $L_b = 0.1$ nH). With the available 6-GHz input (interrogation) power set at -20 dBm and a 10-MHz, -10 dBm data signal present, the conversion loss L_c of the quasi-optical mixer was analyzed for various bowtie antenna lengths. L_c is defined in the transducer sense as the ratio of the third-order response signal power (12 GHz $\pm$ 10 MHz, SSB) dissipated in the bowtie radiation resistance to

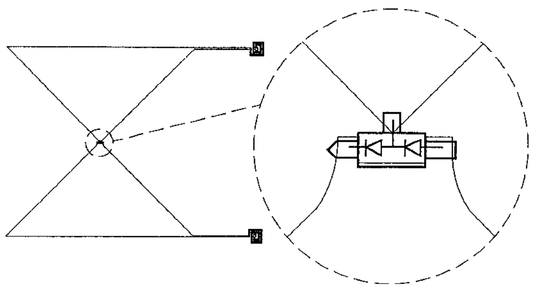


Fig. 6. Layout of the quasi-optical mixer, showing diode mounting and spiral feed-through inductors for data signal.

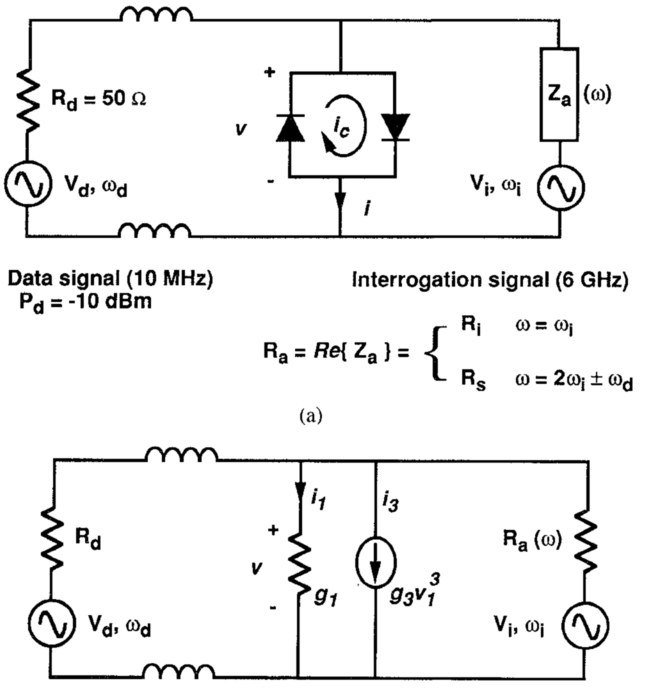


Fig. 7. Equivalent circuits for the quasi-optical mixer: (a) harmonic balance simulation (nonideal diodes); (b) approximate Volterra series analysis.

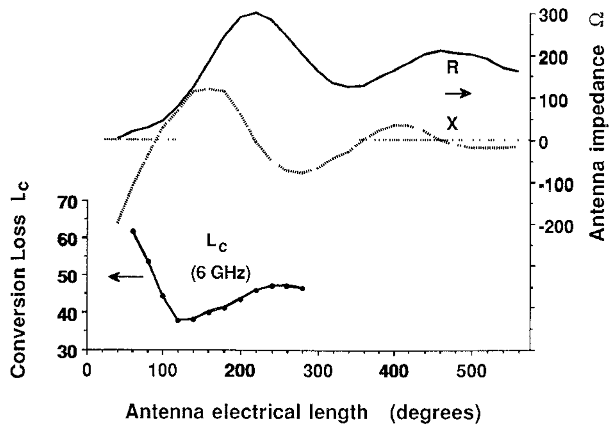


Fig. 8. Mixer conversion loss versus bowtie antenna length (bowtie flare angle = 90°) from harmonic balance circuit analysis.

the interrogation power available from the bowtie. The results indicate an optimum bowtie electrical length of 120–140° at 6 GHz, as shown in Fig. 8. The prototype employed a 200° long bowtie, allowing good performance over the range 4–8 GHz.

B. Mixer Analysis

In normal operation, the signals applied to the quasi-optical mixer are relatively small—the data signal is on the order of -10 dBm, and the received interrogation signal may be significantly smaller for typical interrogation distances. The situation is equivalent to a subharmonically pumped mixer, operating as an upconverter, with a large-signal RF input and small-signal LO. This makes approximate analysis difficult using a simple time-varying circuit model (conversion matrix), as such an approach would predict zero output at the desired third order response frequencies. An alternative approach is to model the mixing operation as a weakly nonlinear process, treating it more like small-signal multitone intermodulation [12]. A Volterra-series approximation to the current flowing in the mixer circuit can then be found, resulting in a closed form expression for the transponder's conversion loss accurate enough for use in system link budget calculations.

Starting with the diode-pair conductance curve expanded as a power series

$$i = g_1 v + g_2 v^2 + g_3 v^3 + g_4 v^4 + g_5 v^5 + \cdots \tag{1}$$

we seek to obtain the third-order current components at the desired response frequencies $2\omega_i \pm \omega_d$ as a function of the two-tone excitation. For the ideal antiparallel diode pair, the well-known conductance equation $i = 2I_s \sinh(\alpha v)$ contains only odd terms, with the coefficients $g_n = 2I_s(\alpha v)^n/n!$. The presence of weak rectified current flowing in the diode loop will affect this small signal characteristic, creating a self-bias effect described in the Appendix. We proceed using the method of nonlinear currents [13], where the terms in (1) are interpreted as current sources with nonlinear dependence on the voltage v across the diode pair (g_1 remains a linear conductance). These are then rearranged such that each source contains only mixing products of the same order, as shown in Fig. 7(b), which makes the equivalent circuit linear. The higher-order currents are still nonlinearly dependent on lower-order voltages, but can be found recursively starting with the first-order voltage. This linear response is found by inspection to be

$$v_1(t) = \frac{1}{2}\left(\frac{1}{R_d g_1 + 1}\right) V_d e^{j\omega_d t} + \frac{1}{2}\left(\frac{1}{R_i g_1 + 1}\right) V_i e^{j\omega_i t} \tag{2}$$

which, along with the complex-conjugate negative frequency component forms a real signal. The equivalent circuit of the mixer can be simplified further by assuming resonance between the diode capacitance and the antenna inductance (approximately valid for the optimum length bowtie), so that the diode-pair embedding impedances at the three frequencies of interest (interrogation R_i, data signal R_d, response R_s) are purely real. Only the data signal sees R_d, as the inductors are essentially open at 6 and 12 GHz. The real third-order current component at the desired response frequency is now readily found

$$i_3(t) = \frac{3}{4} \frac{g_3}{(R_i g_1 + 1)^2 (R_d g_1 + 1)} |V_i^2 V_d| \cos(2\omega_i \pm \omega_d)t. \tag{3}$$

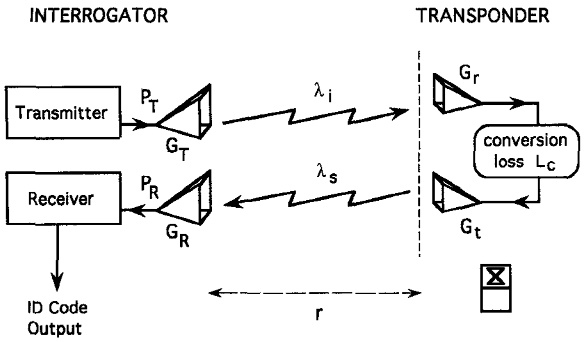


Fig. 9. Interrogator-transponder radio link.

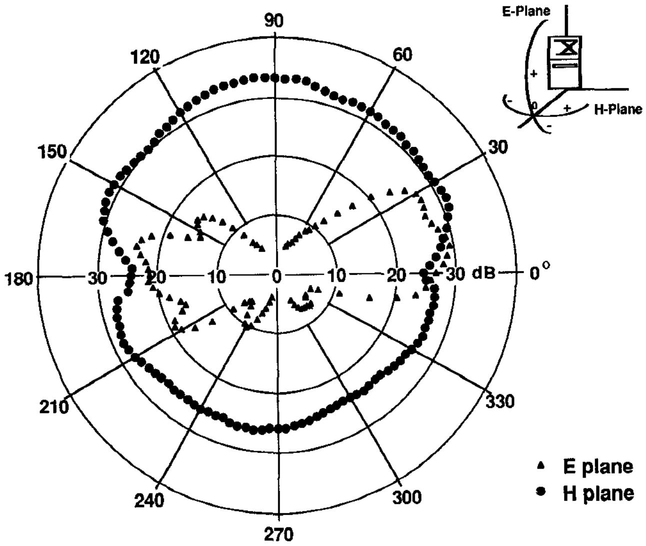


Fig. 10. Conversion radar cross-section (CRCS) of the transponder, E and H planes. Incident intensity is 50 μW/cm^2. 0 dB reference area is −50 dB cm^2.

Flowing in the bowtie antenna, this current produces the radiated response power

$$P_s = \frac{i_3^2 R_s}{2(R_s g_1 + 1)^2}. \tag{4}$$

Expressing the interrogation and data signal voltages in terms of their available powers P_r and P_d, respectively, we obtain the expression for the transponder's conversion loss L_c

$$\begin{aligned} L_c \equiv \frac{P_r}{P_s} &= \frac{(R_d g_1 + 1)^2 (R_i g_1 + 1)^4 (R_s g_1 + 1)^2}{144 \cdot R_d R_i^2 R_s g_3^2 P_d P_r} \\ &= K P_r^{-1}. \end{aligned} \tag{5}$$

Lumping all of the parameters that are fixed for a given transponder (diode parameters, data signal power, circuit impedances) into the constant K shows that, for small signal conditions, conversion loss of the quasi-optical mixer varies inversely with the power received from the interrogation beam. This is a direct consequence of the third-order origins of the response signal.

C. Conversion Loss/RCS

In expressing the conversion loss of the quasi-optical mixer, attention must be paid to the fact that input and output signals are coupled via directive antennas to the free space radiation fields, and as a result no well defined circuit ports exist for power measurements. Incident and scattered field intensities can be measured accurately however, using established radar cross-section (RCS) measurement techniques; it is therefore natural to use a scattering-type metric for evaluating the transponder's performance. A typical RF/ID system radio link is shown in Fig. 9, where the transponder is modeled as receiving and transmitting antennas, with gains G_r and G_t, respectively, connected by a unilateral conversion loss L_c. The scattering properties of the transponder can be expressed by defining a conversion radar cross-section (CRCS) $\tilde{\sigma}$, equivalent to the conventional RCS but with nonequal incident and scattered frequencies, corresponding to the interrogation and response signals. Applying the Friis transmission formula [14] to the link results in the relation

$$\tilde{\sigma} = \frac{\lambda_i^2}{4\pi} \frac{G_r G_t}{L_c}. \tag{6}$$

Although the individual antenna gains and mixer conversion loss cannot be measured separately in an integrated quasi-optical circuit, their ratio in (6) can be measured. In this sense, the CRCS is closely related to the isotropic conversion loss $L_{\rm iso} = L_c/G_r$ defined by Stephan for quasi-optical mixers [9], with the difference that the mixer output (the IF) in the scattering case is not available at a mixer circuit port but rather is re-radiated to the far field with gain G_t. The CRCS is directly measurable, as in Fig. 10, which shows the E and H plane cuts of the transponder's back-scattering pattern. The card has a nearly omnidirectional response in the H plane, which encompasses the most important interrogation angles for a card held upright by a person. Response degrades above, below, and (at longer ranges) directly to the sides of the card due to the pattern nulls of the mixer bowtie and interrogation-detection slot, respectively.

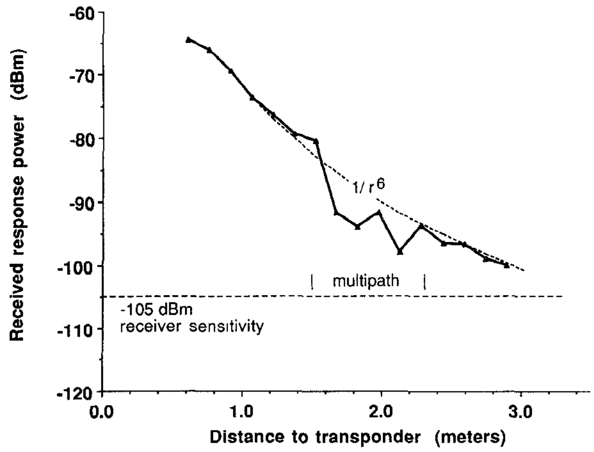


Fig. 11. Response signal power received by interrogator versus distance to transponder.

For link calculations, where the response power back-scattered to the interrogator must be known, the CRCS is used directly in the radar equation. As the transponder's conversion loss L_c was found to vary inversely with received interrogation power, the radar equation takes the interesting form

$$P_R = \frac{(P_T G_T)^2 G_R \lambda_i^4 \lambda_s^2 G_r^2 G_t}{K(4\pi r)^6} \propto \frac{1}{r^6}. \tag{7}$$

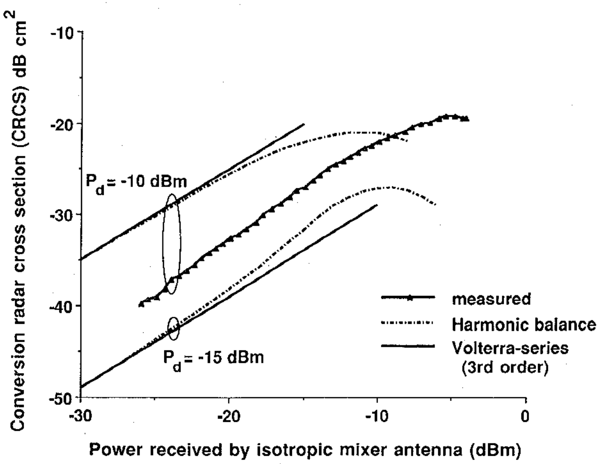


Fig. 12. CRCS of the transponder versus interrogation power received by the mixer antenna.

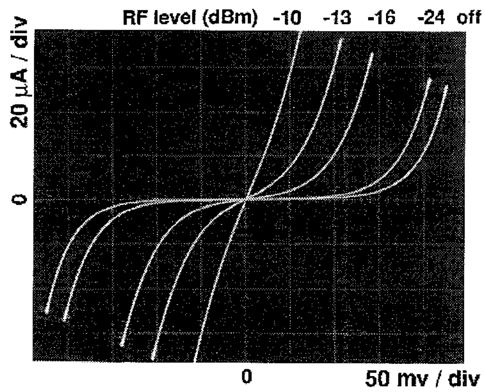


Fig. A1. Measured dc i-v curves of antiparallel diode pair in the bowtie mixer for several incident RF power levels, illustrating effect of self-bias current.

This $1/r^6$ range behavior contrasts with the $1/r^4$ characteristic of a range-independent RCS and was evident in experimental measurements (see Fig. 11). This limits the range of the subharmonically interrogated transponder, although the maximum range using the test system described earlier is 10 ft. (3 m). Increasing the EIRP, or using an optimum-bandwidth receiver could increase this range to over 15 ft. (5 m). For fixed-frequency operation, additional tuning could be added to the mixer to further reduce conversion loss. Fig. 12 shows the variation of CRCS, hence conversion loss, with received interrogation power. Theoretical results from both harmonic balance simulations and the approximate Volterra-series expression are plotted along with experimentally measured data. The offset between the theory and measurement is mainly due to the unrealistic assumption of isotropic transponder antennas in the models. However, if the harmonic balance simulation is assumed accurate, the actual antenna gains can be deduced from the measured data. Since the mixer compression point depends only on the input level, the observed offset in these points on the simulation and measurement curves is attributable directly to transponder receiving gain G_r, which is evidently $-$ 5 dBi at boresight. From (7), the power scattered by the transponder varies with the product $G_r^2 G_t$, therefore the offset in dB between the linear portions of the two curves is equal to $2G_r + G_t$. This indicates that the transponder radiates the 12 GHz-centered response signals with gain $G_t = +\,2$ dBi.

IV. Conclusion

The practicality of an inexpensive microwave identification transponder was demonstrated, emphasizing the advantages of the subharmonic interrogation approach. An approximate, analytical model was derived for the frequency conversion and scattering functions of the quasi-optical circuit, which can be used in system calculations for this and similar transponders. Suggestions were made for future development that could make the transponder a commercially viable identification device.

Appendix

In describing the antiparallel diode pair as a mixing element, it was noted that all even-order currents including dc (rectified RF) flow only in the diode loop and so were neglected in the expression for the conductance seen external to the pair. It was found experimentally that loop dc currents arising from small applied RF signals caused a significant bias offset in the diodes. This is evident in Fig. A1, which is the measured dc i-v curve of the diode pair in the bowtie mixer, while under illumination from a 6-GHz source at various incident intensities (RF power levels are for an isotropic receiving antenna). This can be explained via the expression for diode loop current, $i_c = I_s(\cosh \alpha v - 1)$, [15] which in the presence of an applied RF signal $v = V_L \cos \omega t$ becomes

$$\begin{aligned} i_c &= I_s(\cosh(\alpha V_L \cos \omega t) - 1) \\ &= I_s\{-1 + \boldsymbol{I}_0(\alpha V_L) + 2\boldsymbol{I}_2(\alpha V_L)\cos 2\omega t \\ &\quad + 2\boldsymbol{I}_4(\alpha V_L)\cos 4\omega t + \cdots\} \end{aligned} \tag{A1}$$

where the $\boldsymbol{I}_n$ are the modified Bessel functions of the first kind. The first two terms represent a rectified dc current, the *self-bias* current $I_o = I_s\{\boldsymbol{I}_0(\alpha V_L) - 1\}$. This current flows in the loop, where it forward-biases each diode to a voltage V_o given by the ideal diode equation, $V_o = (1/\alpha)\ln(\boldsymbol{I}_0(\alpha V_L))$. A small signal voltage v applied to the diode pair in its biased state will appear superimposed on the dc bias voltage V_o. The resulting current, which flows in the circuit external to the diodes, is given by

$$\begin{aligned} i &= I_s(e^{\alpha(V_o+v)} - e^{\alpha(V_o-v)}) \\ &= 2I_s e^{\alpha V_o}\sinh \alpha v. \end{aligned} \tag{A2}$$

Inserting V_o yields

$$i = 2I_s \boldsymbol{I}_0(\alpha V_L)\sinh \alpha v \tag{A3}$$

which is the small-signal i-v characteristic of an ideal antiparallel diode pair under self-bias conditions, i.e., in the presence of a larger signal voltage. The voltage V_L represents the larger of the two signals applied to the mixer, which will generally be the transponder's data signal (-10 dBm). This assumes negligible voltage drop across the signal source and diode series resistances, a fair approximation given the small linear conductance of the diode pair. This equation is plotted in Fig. A2 for the same RF illumination as the measured curves, using the -5 dBi estimate obtained earlier for the 6-GHz

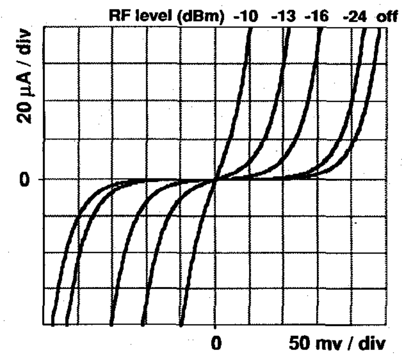


Fig. A2. Theoretical i-v curves (A3) of antiparallel diode pair under same RF illumination as in Fig. A1. A 6-GHz bowtie antenna gain of −5 dBi is assumed.

bowtie gain, showing good agreement. Expanding (A3) in its Taylor series yields the corrected conductance coefficients $g_n = 2I_s(\alpha v)^n I_0(\alpha V_L)/n!$ (n odd). It is interesting to note that use of these coefficients, which are now functions of the applied voltage, in the Volterra analysis gives good agreement with the harmonic balance simulation results in Fig. 12—even with fairly large data signal powers. The nonlinear relation between data signal power and conversion loss is also correctly accounted for, which is missed otherwise [see (5)]. The effects of bias-point offset can thus be included in the Volterra series analysis, which generally does not consider dc current terms, improving the accuracy of the method for larger excitations.

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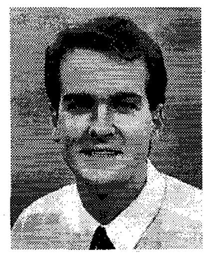
References

[1] D. Mawhinney, "Microwave tag identification systems," *RCA Rev.*, vol. 44, pp. 589–610, Dec. 1983.
[2] "An automatic vehicle ID system for toll collecting," Lawrence Livermore National Laboratory, Rep. No. UCRL-TB-113409, Apr. 1993.
[3] P. de Bruyne and P. Leuthold, "Radar surveillance of Autobahn toll stickers in Switzerland," in *Proc. 1985 Carnahan Conf. on Security Techn.*, Univ. of Kentucky, OES Pub. No. UKY BU137, pp. 109–115, May 1985.
[4] T. Ohta, H. Nakano, and M. Tokuda, "Compact microwave remote recognition system with newly developed SSB modulation," in *IEEE MTT-S Symp. Dig.*, 1990, pp. 957–960.
[5] R. H. McEachern, K. J. Prost, D. Hampel, and D. D. Mawhinney, "A personnel-tracking system," *RCA Engineer*, vol. 28-6, pp. 57–63, Nov./Dec. 1983.
[6] R. Page, "A low power RF ID transponder," *RF Design*, pp. 31–36, July 1993.
[7] M. Bouthinon, J. Gavan, and F. Zadworny, "Passive microwave transponder, frequency doubler for detecting avalanche victims," in *Proc. 10th European Microwave Conf.*, Sept. 1980, pp. 579–583.
[8] G. R. Sloan, "The modeling, analysis, and design of filter-based parametric frequency dividers," *IEEE Trans. Microwave Theory Tech.*, vol. 41, no. 2, pp. 224–228, Feb. 1993.
[9] K. Stephan and T. Itoh, "A planar quasi-optical subharmonically pumped mixer characterized by isotropic conversion loss," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, pp. 97–102, Jan. 1984.
[10] ——, "Inexpensive short-range microwave telemetry transponder," *Electron. Lett.*, vol. 21, pp. 877–878, Oct. 11, 1984.
[11] G. H. Brown and O. M. Woodward, "Experimentally determined radiation characteristics of conical and triangular antennas," *RCA Rev.*, vol. 13, pp. 425–452, Dec. 1952.
[12] S. A. Maas, *Nonlinear Microwave Circuits.* Norwood MA: Artech House, 1988.
[13] J. J. Bussgang, L. Ehrman, and J. W. Graham, "Analysis of nonlinear systems with multiple inputs," *Proc. IEEE*, vol. 62, no. 8, pp. 1088–1119, Aug. 1974.
[14] H. T. Friis, "A note on a simple transmission formula," in *Proc. IRE*, vol. 34, pp. 254–256, 1946.
[15] M. Cohn, J. E. Degenford, and B. A. Newman, "Harmonic mixing with an antiparallel diode pair," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-23, no. 8, pp. 667–673, Aug. 1975.



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