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Concept of Beam Steerable Transponder based on Load Modulation

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***Abstract*—The concept of beam steering transponder based on load modulation is proposed to enhance the received power of backscattered communication devices. The transponder consists of antenna array elements of equal length operates at the Wi-Fi frequency band of 2.4 GHz. By properly weighting the modulated signal in each port, the received power can be maximized in a particular direction depending on the direction of arrival (DoA) of the signal. The weighting is done differently for each direction. It provides up to approximately 11 dB improvement in received power with weighting compared to backscatter communication with a single antenna transponder. The concept is studied theoretically and by simulations.**

***Index Terms*—Antenna array, backscatter, beam steering, modulation, transponder, IoTs, WSNs.**

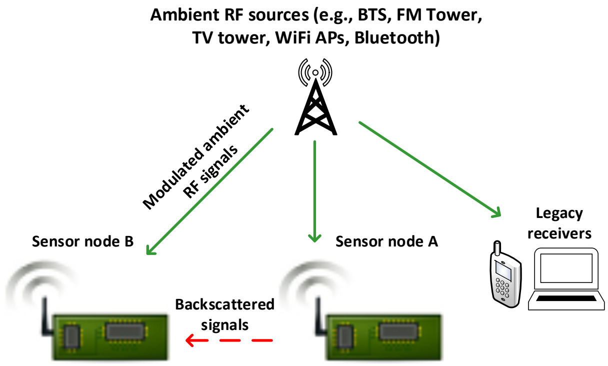

Fig. 1. Ambient backscatter communication systems.

I. INTRODUCTION

Ambient backscatter communication has a great potential for future low-energy communication systems, especially Internet-of-Things (IoT) [1]. One of the most important aspect of widely deploying the IoT sensors in backscatter communication is to reuse the prevalent ambient signals such as Wi-Fi APs, Bluetooth, TV towers, cellular base stations to piggyback their data. If practicable this would eliminate the need for dedicated reader devices for RFID and backscatter communication [2]. The another advantage of such system is that they do not need a dedicated frequency spectrum, which is scarce and costly. Fig. 1 illustrates the example of Ambient Backscatter Communications Systems (ABCSs).

The RF sources in ABCSs are uncontrollable, e.g., transmission power and locations, thus the onus comes on the design of transponder to achieve optimal performance in terms of sufficient throughput and range. In recent times there has been tremendous growth in the designing of IoT devices such as gadgets, smart wearables, fitness trackers, security cameras, temperature sensors etc which can sense, collect, and upload physiological data in a 24x7 manner [3].

The demand to significantly improve the range of backscatter communication in smart devices requires new techniques. In order to use the available RF sources in effective manner, the transponder needs to tune its operation where the highest ambient power is available. The information can be communicated back to the receiver or reader which can be located at different locations. This has led to the development of beam steerable transponder. In our proposed approach, we can have a beam steerable transponder by changing the phase of the modulated signal in each element. The principle is same as in steerable phased arrays but here the difference is that now we first receive the signal from one direction and then direct the modulated signal to the same or another direction depending on the receiver location. To the authors' knowledge, this is the first time when such a technique is proposed for devices based on back- or re-scattering communication principle.

The measurement results in [4] shows that it provides very small increment in received power in a particular direction. It also use circulators and RF switches which are expensive and introduce complexity and losses in the system. Another approach in [5] use analog circuits for load modulation to create a directional backscatter tag which adds complexity. We demonstrate a load modulation technique for backscatter communication by using antenna array elements and maximize the received power in a particular direction and at a given frequency by properly weighting the modulated signals.

We exploit the digital part of the transceiver (IC technology) which advance fast as compared to analog components. The weighting of the signal can be done by using low power microcontroller or some cheap IC. The weights can be calculated mathematically from the scattering parameters of the antenna. The weighting techniques in [6] for the antenna feeding signals have been used to improve antenna efficiency in a coupled environment but it is not applicable to transponder case. In our work we propose a new theoretical approach to validate our concept. To the best of authors' knowledge, such transponders, have not been presented previously. The suitability of the proposed approach is demonstrated theoretically and by simulations.

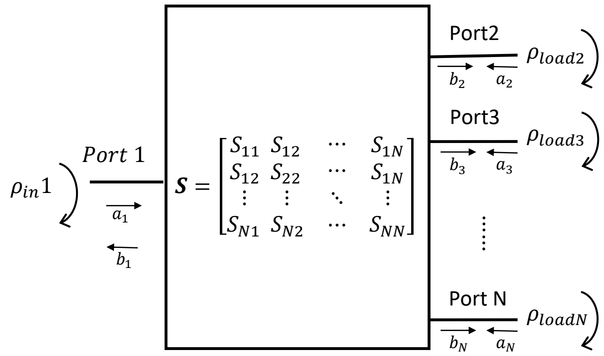

Fig. 2. S-parameter block with switchable loads.

II. Theory

A. Operational principle

Fig. 2 illustrates the operation principle of the studied transponder where port 1 represents the reader antenna and ports 2, 3, 4 and 5 depict the four antenna elements connected to switchable loads. Ports 2 to 5 are switched between two loads, a short circuit (zero impedance) and an open circuit (infinite impedance) using a switch-control signal which is a periodic square wave signal of period T = $1/f_m$. The Fourier series of a square wave has odd harmonics. This means a square wave of period $1/f_m$ has sinusoidal signals of frequencies f_m, $3f_m$, $5f_m$, ... [7]. A signal of frequency f_1 ($f_1 \gg f_m$) at the switch, gets mixed with the harmonics of the square wave resulting in the modulated signal with frequencies $f_1 \pm f_m$, $f_1 \pm 3f_m$, $f_1 \pm 5f_m$, and so on.

B. Modulated reflection coefficient

Since the load switches between a short and an open circuit (implying reflection coefficients -1 and $+1$ respectively), the modulated signal is reflected at the input of the switch. Switching the load between open and short circuits cause time-domain reflection coefficient $\rho(t)$ at the input of the switch, that varies between $\rho_1 = -1$ and $\rho_2 = +1$ respectively. The modulated reflection coefficient at the input of the switch due to the changing loads in the frequency domain $\rho_{load}(\omega_m)$ = ρ_{load} is the Fourier transform of the time-domain reflection coefficient $\rho(t)$ over one period T. This is calculated as:

$$\begin{aligned} \rho_{load}(\omega_m) &= \frac{1}{T}\int_0^T \rho(t)\ e^{-j\omega_m t} dt \\ &= \frac{\omega_m}{2\pi}\left(\int_d^{d+\pi/\omega_m} \rho_1\ e^{-j\omega_m t} dt \right. \\ &\quad \left. + \int_{d+\pi/\omega_m}^{d+2\pi/\omega_m} \rho_2\ e^{-j\omega_m t} dt\right). \end{aligned}$$

This becomes

$$\rho_{load} = \frac{-j}{2\pi}\ e^{-2jd\omega_m}(e^{jd\omega_m}(2\rho_1 - \rho_2) - \rho_2) \tag{1}$$

where, $\omega_m = 2\pi f_m$ and d is the delay time from where a period of the square wave begins. When $d = 0$,

$$\rho_{load} = \frac{-j}{\pi}(\rho_1 - \rho_2)\ , \tag{2}$$

(1) gives the modulated reflection coefficient ρ_{load} at the first harmonic. The performance of the transponder in this work is studied at the first harmonic only hence, (1) suffices to model the transponder. Note that in (2) the amplitude of ρ_{load} is directly proportional to the difference between ρ_1 and ρ_2, and is independent of ω_m when $d = 0$.

The studied transponder is modeled using multiple modulators. The output of the modulators are connected to a multi-element array antennas of equal length. To create the phase difference between the output of the modulators, the switch-control signals of these modulators are delayed in the time domain. This is done because a delay in the time domain corresponds to a phase difference in the frequency domain.

C. Derivation of general solution of N-port network based on S-parameters

In this section, the input reflection coefficient of an N-port microwave network is studied and the received modulated power based on S-parameters is calculated. Consider an N-port network as shown in Fig. 2 with reflection coefficients ρ_{loadj} at the loads, input reflection coefficient ρ_{in1} and the scattering parameters S_{ij}. The following derivation is applied from [8].

The input reflection coefficient at Port 1 is:

$$\rho_{in1} = \frac{b_1}{a_1}, \tag{3}$$

and the load reflection coefficients at Ports $2, 3, ..., N$ are:

$$\rho_{loadn} = \frac{a_n}{b_n} \text{ where, n= 2,3,...,N.} \tag{4}$$

From the definition of scattering matrix

$$[\mathbf{b}] = \begin{bmatrix} b_1 \\ b_2 \\ \vdots \\ b_N \end{bmatrix} = [\mathbf{S}][\mathbf{a}] = \begin{bmatrix} S_{11} & S_{12} & \dots & S_{1N} \\ S_{21} & S_{22} & \dots & S_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ S_{N1} & S_{N2} & \dots & S_{NN} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ \vdots \\ a_N \end{bmatrix} \tag{5}$$

Using (3) and (4), (5) can be rewritten as:

$$[\mathbf{b}] = \begin{bmatrix} b_1 \\ b_2 \\ \vdots \\ b_N \end{bmatrix} = [\mathbf{S}][\mathbf{a}] = \begin{bmatrix} S_{11} & S_{12} & \dots & S_{1N} \\ S_{21} & S_{22} & \dots & S_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ S_{N1} & S_{N2} & \dots & S_{NN} \end{bmatrix} \begin{bmatrix} b_1/\rho_{in1} \\ b_2\rho_{load2} \\ \vdots \\ b_N\rho_{loadN} \end{bmatrix}$$

$$= \begin{bmatrix} S_{11} & S_{12} & \dots & S_{1N} \\ S_{21} & S_{22} & \dots & S_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ S_{N1} & S_{N2} & \dots & S_{NN} \end{bmatrix} \begin{bmatrix} 1/\rho_{in1} & 0 & \dots & 0 \\ 0 & \rho_{load2} & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & \rho_{loadN} \end{bmatrix} \begin{bmatrix} b_1 \\ b_2 \\ \vdots \\ b_N \end{bmatrix}$$

or,

$$[\mathbf{b}] = \begin{bmatrix} S_{11}/\rho_{in1} & S_{12}\rho_{load2} & \dots & S_{1N}\rho_{loadN} \\ S_{21}/\rho_{in1} & S_{22}\rho_{load2} & \dots & S_{2N}\rho_{loadN} \\ \vdots & \vdots & \ddots & \vdots \\ S_{N1}/\rho_{in1} & S_{N2}\rho_{load2} & \dots & S_{NN}\rho_{loadN} \end{bmatrix} \begin{bmatrix} b_1 \\ b_2 \\ \vdots \\ b_N \end{bmatrix} = [\mathbf{A}][\mathbf{b}]$$

or,

$$\textbf{(A-I)b} = \mathbf{0}, \tag{6}$$

where $\mathbf{I}$ is the identity matrix. ρ_{in1} in matrix $\mathbf{A}$ is unknown whereas the S-parameters and ρ_{loadj} are known.
(6) has two solutions. The first is the zero column vector $\mathbf{b} = 0$, implying that the input port is matched to the impedance of the source and all other ports are matched to the load impedances. For a transponder to function, we must have $\mathbf{b} \neq \mathbf{0}$ (port 1 is neither perfectly matched nor isolated). For a solution of (6) to exist in this case, the determinant of **(A-I)** must be zero.

$$\textbf{det(A-I)} = \begin{vmatrix} S_{11}/\rho_{in1} - 1 & S_{12}\rho_{load2} & \dots & S_{1N}\rho_{loadN} \\ S_{21}/\rho_{in1} & S_{22}\rho_{load2} - 1 & \dots & S_{2N}\rho_{loadN} \\ \vdots & \vdots & \ddots & \vdots \\ S_{N1}/\rho_{in1} & S_{N2}\rho_{load2} & \dots & S_{NN}\rho_{loadN} - 1 \end{vmatrix} = 0$$

Using the square matrix properties, the determinant of a square matrix is equal to the determinant of its conjugate transpose or Hermitian [8].

$$\textbf{det(A-I)} = \textbf{det(A-I)}^H = \mathbf{0}, \tag{7}$$

or,

$$\begin{vmatrix} S_{11}/\rho_{in1} - 1 & S_{21}/\rho_{in1} & \dots & S_{N1}/\rho_{in1} \\ S_{12}\rho_{load2} & S_{22}\rho_{load2} - 1 & \dots & S_{N2}\rho_{load2} \\ \vdots & \vdots & \ddots & \vdots \\ S_{1N}\rho_{loadN} & S_{2N}\rho_{loadN} & \dots & S_{NN}\rho_{loadN} - 1 \end{vmatrix} = 0 \tag{8}$$

Solving (8) gives,

$$\rho_{in1} = \frac{S_{11}D_1 - S_{21}D_2 + S_{31}D_3 - ..... \pm S_{N1}D_N}{D_1} \tag{9}$$

(9) gives solution for an odd and even numbers of N, where D_1, D_2 and D_3 are the determinants of the minors, obtained of the matrix $\textbf{(A-I)}^H$ by removing the first row and first, second, and third column of the matrix and so on. It is evident from (9), that ρ_{in1} is S_{11} modified by a term determined by the the S-parameters (excluding S_{11}) and the load reflection coefficients. It also shows how the input reflection coefficient is governed by the load reflection coefficients of the multi antenna elements at the transponder. When the scattering matrix $\mathbf{S}$ and the load reflection coefficients ρ_{loadj} are known, the input reflection coefficient ρ_{in1} can be solved from (9).

D. Modulated received power based on S-parameters

The input reflection coefficient ρ_{in1} is the ratio of the reflected voltage wave to the incident voltage wave. Therefore, a modulated reflection coefficient $\delta\rho_{in1}$ would be defined as the ratio of the backscattered modulated voltage waves at modulated frequency $f_1 + f_m$ to the transmitted voltage wave at frequency f_1. This means that a reader transmitting 0 dBm of power receives $\delta\rho_{in1}$ dBm of power at the modulated frequency from the transponder.

In the dBs,

$$\textit{Received power at } f_1 + f_m = \delta\rho_{in1} + \textit{transmit power}. \tag{10}$$

The input reflection coefficient ρ_{in1} of (9) has components due to the modulated signal and S_{11}. S_{11} is the reflection coefficient of the reader antenna, and it must be subtracted from the input reflection coefficient to get the modulated reflection coefficient. Therefore, the modulated reflection coefficient ($\delta\rho_{in1}$) at port 1, when transponder have four antenna elements i.e., N = 5 is:

$$\delta\rho_{in1} = \rho_{in1} - S_{11} = \frac{-S_{21}D_2 + S_{31}D_3 - ..... + S_{51}D_5}{D_1} \tag{11}$$

Similarly, the modulated reflection coefficient $\delta\rho_{in1}$ of the reference transponder with a single antenna element can be modeled as a 2-port network. Thus,

$$\rho_{in1} = S_{11} + \frac{S_{12}S_{21}\rho_{load2}}{1 - \rho_{load2}S_{22}},$$

and

$$\delta\rho_{in1} = \frac{S_{12}S_{21}\rho_{load2}}{1 - \rho_{load2}S_{22}}. \tag{12}$$

E. Maximizing the received power with optimal delays

The goal of our work is to study whether the phases i.e., time domain delays, of the modulators of the studied transponder can be optimized such that its backscattered signal power is maximized. From (1) and (11), we can see that the modulated reflection coefficient $\delta\rho_{in1}$ is a function of the time-domain delay d of the switch-control signal at the different modulators. Therefore, using delays in the switch-control signals give phase-variable modulators. The switch control signal of the modulator at port 2 is used as a reference, and the switch control-signal of the modulators at port 3, 4 and 5 are delayed with respect to the reference. Thus, delay $d = 0$ for ρ_{load2}, and the optimal delay values $d3$, $d4$ and $d5$ for ρ_{load3}, ρ_{load4} and ρ_{load5}, respectively, are used to maximize the modulated reflection coefficient.

The modulated reflection coefficients for the studied and reference transponders were calculated using (11) and (12), respectively. The delay values of (11) are optimized numerically to give largest modulated reflection coefficient $\delta\rho_{in1}$ for a given frequency which is 2.4 GHz in our case [10]. One set of optimal delays are only valid for a given frequency and we calculate the optimal delay values at each direction to enhance the received power in that direction. The following section demonstrate the simulations based on the theoretical approach presented in this section.

III. SIMULATIONS

A. *Transponder design*

The transponder is intended to operate at the Wi-Fi frequency band of 2.4 GHz. The reason for choosing this frequency is that the Wi-Fi emitters as a ambient RF source are available everywhere in the surroundings. The antenna model used for the studied transponder, shown in Fig. 3(a), consists of four monopole antenna elements of equal length. The length of the monopole is 28 mm resonating at the frequency of 2.4 GHz (S_{22}, S_{33}, S_{44}, S_{55} < -10 dB). The optimal number of antenna elements in the transponder is something to take under consideration in future studies. Less elements reduce the cost, size and complexity of the studied transponder. The diameter of the monopoles is 2 mm and the spacing between the antenna elements is kept as $\lambda/4$. Different spacing between the elements can also be selected for transponder design and we can not say it is optimal selection. The thickness of the ground plane is 2.5 mm. The antennas are fed through 50Ω coaxial lines (SMA connector)

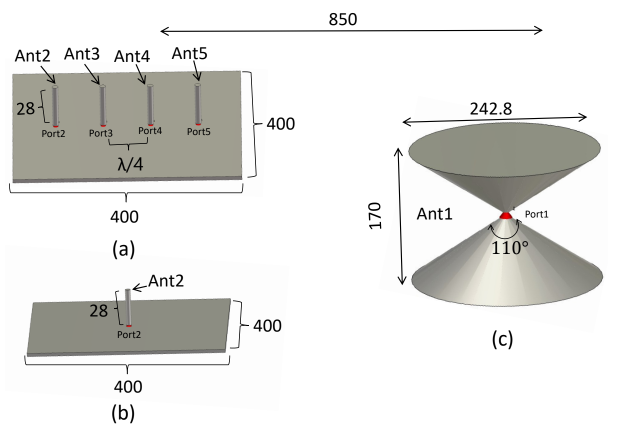

Fig. 3. (a) Studied transponder model with four monopoles (b) Reference transponder model with single monopole, (c) Reader antenna model. All dimensions are in millimeters. The reader and transponder antennas are not to scale.

The reader antenna in Fig. 3(c) is a finite biconical antenna [9]. The biconical antenna has a very large bandwidth, and it can illuminate the transponder antennas in the desired frequency of 2.4 GHz. Fig. 4 shows the realized gain of reader antenna. The distance between the transponder and the reader antenna is 850 mm which is the farfield distance. Both of these antenna models are modelled in CST Microwave Studio, and their scattering matrices are exported from CST to model the transponder. The reference transponder shown in Fig. 3(b) is used as a reference, and its performance is compared with the studied transponder. The antenna model used for the reference transponder contains a single element monopole antenna, which is tuned at 2.4 GHz.

B. *Simulation results*

Fig. 5 depicts general illustration how simulations are performed at different angles to analyze the directional pattern of proposed transponder. A traditional backscatter device reflects

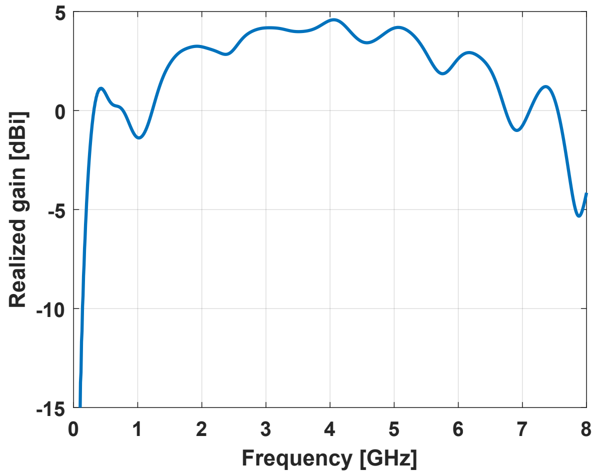

Fig. 4. The realized gain of the reader antenna to the direction of the transponder.

the signal back in all directions, but the signal of interest is only in the direction of the reader. The reader device (ambient source) can be in any direction and in our proposed approach, we can steer the beam in any particular direction and enhance the received power based on optimal delays. The optimal delay values are optimized at each angle to maximize the received power.

The result of the simulations, shown in Fig. 6, shows the received power at frequency $f_1 + f_m$ at the reader antenna when the transmitted power at frequency f_1 (2.4 GHz) is 0 dBm. Fig. 6 is the main result of our work and it also shows the comparison with reference transponder. The received power improves by 8 dB in end fire array direction (0°) as compared to no delays and up to 11 dB in comparison to reference transponder. This about 8 dB improvement is also in other directions except in broadside array direction (90°)

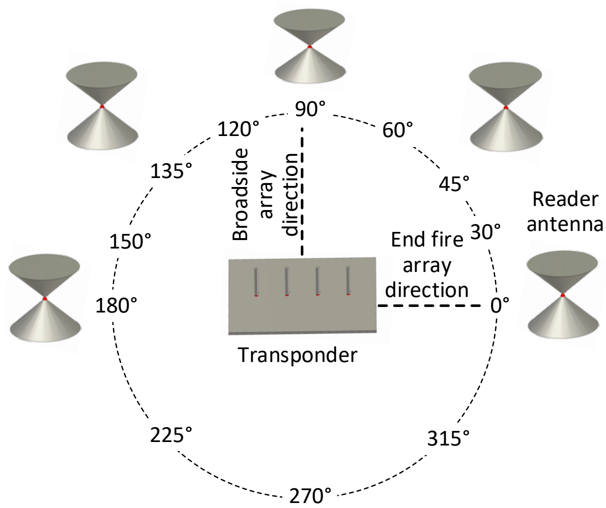

Fig. 5. Illustration of simulations performed in CST at different angles to investigate the transponder directional pattern

and near angles (80° and (100°). In broadside direction (90°) the received power value with optimal delays and no delays is equal because the transponder is in opposite direction with respect to reader antenna and optimal delays enhance the received power constructively in this direction, but it performs better than the reference transponder up to 4 to 6.5 dB in these directions (80°, 90°, 100°). The received power values from 180° to 360° is not shown due to symmetry.

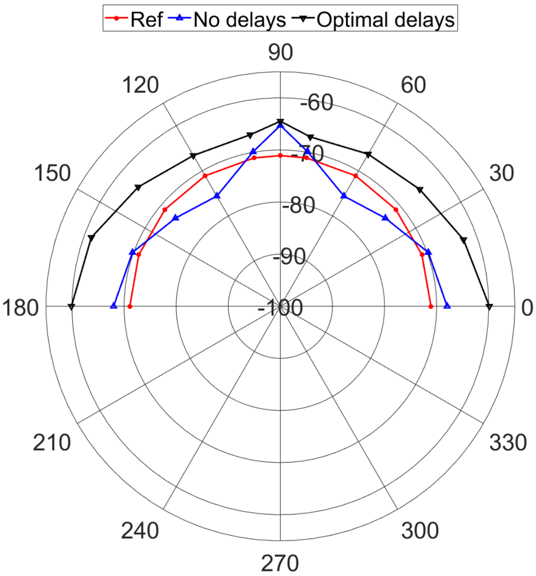

Fig. 6. The modulated received power with no delays and optimal delays of the studied transponder at different angles and its comparison with reference transponder.

IV. Conclusion

In this paper, we have proposed the use of load modulation technique by weighting the modulated signal to significantly improve the received power of backscatter communication. The simulated transponder is a proof of concept to demonstrate the technique, showing promising results. Future work includes designing a transponder prototype to validate our theoretical and simulation results. More research also needed for the practical realization of the weighting, which includes designing of variable-gain amplifiers and tunable phase shifters. Furthermore, to implement the optimal delays digitally with a low power microcontroller or with a cheap IC chip. Different antenna structures should also be investigated. Beam steerable transponder is a needed feature in the ever-growing market of backscatter communication devices.

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