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Contents

List of Figures

Abstract

Sensor nodes distributed over a large area can be used to collect environmental information such as chemical pollution. The sensed information can be communicated to a central location, without significant power consumption by the sensor nodes, by utilizing backscatter modulation. In this approach, a central source such as a radar illuminates the sensor nodes by microwave radiation. The reradiation from each sensor can be modulated by switching a load on the sensor antenna. This permits design of low-power sensors having a long battery life. To avoid interference at the radar from unwanted objects, harmonic reradiation can be used on the sensor nodes and the reradiated harmonic can be processed.

A passive harmonic reradiator is designed to reradiate an incident 10 GHz electromagnetic wave at 20 GHz. The design is for low incident power density. The design comprises of receiving and transmitting integrated patch antennas, a low-barrier Schottky diode and double stub impedance matching. Measurement of the reradiated power, with a representatively input low power density, showed promising characteristics for the use of harmonic reradiator as described above.

1 Introduction

Sensor nodes distributed over a large area can be used to collect environmental information such as chemical or nuclear pollution, or for habitat monitoring. It is desirable that these sensors have a long battery life either because of the large number of sensors typically involved or the inaccessibility of terrain, or both. Active transmission of the sensed information to a central location consumes a significant amount of energy on the sensor nodes and reduces their battery life. In this respect, backscatter modulation mechanism provides an energy efficient way of communicating information.

In the backscatter modulation approach, a central source with adequate energy resources illuminates the sensor nodes. The sensors encode the information by modulating the reradiation. The reradiation can be processed at the central location to decode the transmitted information. Backscatter modulation can be accomplished by switching a load on the sensor antenna, a process which requires very little power consumption. This permits design of sensors having a long battery life.

Microwave radars can illuminate a large number of sensors from a large stand-off distance. Current microwave technology permits significant transmit power levels on the radar. A radar can thus be used as a central source of electromagnetic radiation and to collect information from the sensor nodes. In such a case, as with any backscatter modulation, the return from unwanted objects interferes with the return from the sensor nodes. This interference can be avoided by utilizing a harmonic reradiator on the sensor nodes and processing the reradiated harmonic at the radar. Backscatter modulation in its simplest form can be On-Off keying of transmitted radar pulses. The harmonic reradiation takes place in the On-state and not in the other.

With this motivation, a passive second harmonic reradiator which reradiates an incident 10 GHz wave at 20 GHz is designed and fabricated. The harmonic reradiator consists of receiving and transmitting integrated patch antennas connected by a low-barrier Schottky diode and a double stub impedance matching network for each antenna (Fig. 1.). The entire design is small and low-cost. Our goal is to design and operate the harmonic reradiator at representatively low incident power densities and study its harmonic reradiation level. This can be used to access if adequate signal to noise ratio can be achieved at the radar for significant operating ranges. For the purpose of calculating the energy available at the radar, the harmonic reradiator can be characterized by a scattering cross-section. For any given power density, the product of the scattering cross-section and the power density is the power radiated by the sensor. Radar equation can be used to calculate the received power. The scattering cross-section is calculated using the measured reradiated levels from the fabricated harmonic reradiator. The scattering cross-sections at low power densities show significant promise for the use of the harmonic reradiator in a low power distributed sensor system.

Active frequency doublers having a power supply have been studied in [1],[2]. Broadband Schottky diode resistive frequency doubler designs have been presented in [3],[4], and passive frequency doublers using Schottky barrier varactor diodes in a different frequency regime

Figure 1: Harmonic reradiator concept

are presented in [5],[6]. However, the input power level in these designs place the sinusoidal voltage at the diode above threshold. In the remove sensor network we envision, the available power to the reradiator can be as low as -10 dBm, and this places the voltage well below the threshold for a Schottky diode.

Our design approach is as follows. We first choose a diode based on a first order evaluation of harmonic reradiation using a simple circuit model. This is described in Section 2. Two patch antennas, one receiving at 10 GHz and the other reradiating at 20 GHz are designed independently. The design is discussed in Section 3. The next step is the design of matching network connecting the diode and the antennas to form a complete harmonic reradiator. This is presented in Section 4. Experimental results are presented in Section 5. In Section 6, an example energy budget is presented, showing the utility of the harmonic reradiator.

2 Diode selection

The diode selection should be such that the behavior is nonlinear at low power levels. The prediction of the extent to which a diode can generate a harmonic at low power levels is difficult owing to the dependence of non-linearity on the power level, presence of both harmonic and fundamental frequencies in the diode voltage and the effect of terminations.

Figure 2: Diode Model

Let us consider a fairly simple nonlinear diode model shown in Fig. 2. The model consists of a variable resistance in parallel with a junction capacitance. For any given voltage, the

currents in the resistor and the capacitor are given by

$$I_R = I_0 \left[\exp \left(\frac{V}{\eta k T} \right) - 1 \right] \tag{1}$$

and

$$I_C = C \frac{\mathrm{d}V}{\mathrm{d}t} + V \frac{\mathrm{d}C}{\mathrm{d}t}, \tag{2}$$

where

$$C = C_0 \frac{1}{\left[1 - \frac{V}{V_T} \right]^{1/2}}. \tag{3}$$

It is desired that the diode be connected in series between two antennas for ease of fabrication. When the diode is connected in series with antennas, the relative non-linearities of the diode nonlinear resistance and the junction capacitance are likely to affect the performance. We first compare the harmonic currents generated in the resistance and capacitance by applying a sinusoidal voltage of 10 GHz across the diode. Let $V = V_0 \cos \omega t$ be the voltage across the diode. The nonlinearity in the current depends on the magnitude of V_0. We use Fourier analysis to determine the percentage of second harmonic in the resistive and capacitive branches. If we choose $V_0 = 0.1$ V, the normalized resistive and capacitive currents obtained are shown in Table 1.

	DC current	Current at 10 GHz	Current at 20 GHz
Resistive branch	0.524	1	0.626
Capacitive branch	0	1	0.173

Table 1: Normalized currents due to resistive and capacitive non-linearities

From the Fourier analysis, we observe that the junction resistance has much larger nonlinearity than the junction capacitance. Further, the behavior is similar for different magnitudes of the applied voltage. When we try to select a diode to generate harmonics, one criteria is to select a diode whose junction resistance is much smaller than the impedance due to junction capacitance, to let a greater percentage of current flow through the junction resistance.

Based on this study, the Agilent HSCH-9161 GaAs Schottky diode is selected. The diode has a junction capacitance of 0.035 pF, a saturation current of about 12×10^{-6} A, and a series resistance of 50 Ω. This diode has the smallest ratio of the junction resistance impedance to the junction capacitance impedance that we found on the market.

3 Antenna design

To receive the electromagnetic wave at the fundamental and to radiate the harmonic, a good antenna design is needed. A dual-band antenna for the fundamental and second harmonic

(a) Top view

(b) Side view

Figure 3: Microstrip patch antenna.

frequencies can achieve the functions we need. However, the dual-band antenna with only one port is difficult to integrate with the non-linear load, and the design of matching network is complicated. Hence, one antenna at the fundamental frequency and another at the second harmonic are designed. The diode and the matching networks are connected between the two antennas.

Some of the considerations for the antenna design for this specific application are a) low cost, b) low directivity (to ensure that sensors need not be pointed in any particular direction to receive the signal) and c) planar (for ease of use). Microstrip antennas form an ideal choice for these applications. Microstrip patch antennas have good performance and have been investigated for many years [7]. Rectangular patch antennas are chosen since they can be fabricated at a low cost.

Design of patch antennas

The geometry of a patch antenna is shown in Fig. 3. The antenna design involves a choice of the substrate material, choice of the feed method, and the calculation of the dimensions of the antenna.

Choice of substrate

Patch antennas having a thick substrate with a low dielectric constant have a good efficiency and a large bandwidth. However, increasing the substrate thickness also enlarges the surface waves and spurious feed radiation. So, the substrate cannot be too thick [8]. Based on our simulations and experience we use Duroid 5880 as the substrate material. This substrate material has a low dielectric constant of 2.2 and a thickness of 0.508 mm and can operate at high frequencies.

Antenna Dimensions

To calculate the dimensions of the antenna for the desired resonant frequency, the transmission line model developed in [7] is used. Though the transmission line model is not very accurate, this is used to obtain an initial estimate of the dimensions. The dimensions are fine tuned based on simulations using the moment method.

The effective length of the antenna $L_{eff} = L + 2\triangle L$ determines the resonant frequency f_r of the antenna. $\triangle L$ accounts for the edge effects. The resonant frequency is given by

$$f_r = \frac{1}{2L_{eff}\sqrt{\epsilon_{reff}}\sqrt{\mu_0\epsilon_0}} \tag{4}$$

$$= \frac{v_0}{2L_{eff}\sqrt{\epsilon_{reff}}}, \tag{5}$$

where ϵ_{reff} accounts for fringing effects and v_0 is the speed of light in free space. A practical width W, which leads to good radiation efficiencies is given by [7]

$$W = \frac{v_0}{2f_r}\sqrt{\frac{2}{\epsilon_r + 1}}. \tag{6}$$

The physical length L of the strip is given by

$$L = \frac{v_0}{2f_r\sqrt{\epsilon_{reff}}} - 2\triangle L. \tag{7}$$

ϵ_{reff} and $\triangle L$ can be computed using the following equations. It is assumed that $W/h > 1$.

$$\epsilon_{reff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2}\sqrt{(1 + 12\frac{h}{W})}. \tag{8}$$

$$\frac{\triangle L}{h} = 0.412\left[\frac{(\epsilon_{reff} + 0.3)\left(\frac{W}{h} + 0.264\right)}{(\epsilon_{reff} - 0.258)\left(\frac{W}{h} + 0.8\right)}\right]. \tag{9}$$

Antenna feed

Several designs for feeding the antenna exist. However, many of them require high resolution in fabrication, especially when the operating frequency is high. The direct feed line is chosen to feed the edge of the patch antenna as such a structure is easier to fabricate [9],[10].

The direct feed line is a microstrip transmission line connected to the patch antenna. For a given substrate, the characteristic impedance of the transmission line is determined by its width W_0 (Fig. 3(a)). The characteristic impedance is given by [11]

$$Z_0 = \begin{cases} \dfrac{60}{\sqrt{\epsilon_{reff}}} \ln(\dfrac{8h}{W_0} + \dfrac{W_0}{4h}), & \dfrac{W_0}{h} \leq 1 \\ \dfrac{120\pi}{\sqrt{\epsilon_{reff}}(\frac{W_0}{h} + 1.393 + 0.667 \ln(\frac{W_0}{h} + 1.444))}, & \dfrac{W_0}{h} > 1. \end{cases} \tag{10}$$

Our design uses a characteristic impedance of 50 Ω. The width of the transmission line for a given characteristic impedance can be computed by using

$$\frac{W_0}{h} = \begin{cases} \dfrac{8e^A}{e^{2A} - 2}, & \dfrac{W_0}{h} < 2, \\ \dfrac{2}{\pi}(B - 1 - \ln(2B - 1) + \dfrac{\epsilon_r - 1}{2\epsilon_r}(\ln(B - 1) + 0.39 - \dfrac{0.61}{\epsilon_r}), & \dfrac{W_0}{h} > 2, \end{cases} \tag{11}$$

where

$$A = \frac{Z_0}{60}\sqrt{\frac{\epsilon_r + 1}{2}} + \frac{\epsilon_r - 1}{\epsilon_r + 1}(0.23 + \frac{0.11}{\epsilon_r}), \quad \text{and} \tag{12}$$

$$B = \frac{377\pi}{2Z_0\sqrt{\epsilon_r}}. \tag{13}$$

The position d of the transmission line determines the impedance seen at the feed-point. The position d for an input impedance of 50 Ω is determined using simulations.

Performance simulations for 10 GHz and 20 GHz antennas

The dimensions of the antennas can be calculated using the formulas presented in the Sec 3. The resulting dimensions are shown in Table 2. Using these parameters of the patch antennas as the initial parameters in the Momentum simulation of Agilent Advanced Design System, we simulate the patch antennas and optimize the parameters of the patch antennas. The parameters of the structures and simulation results are shown in Fig. 4 - 5 and Table 3.

4 Impedance matching

The matching circuits are designed assuming that each antenna presents an impedance of 50 Ω at its operating frequency. Since the antenna impedance at the harmonic frequencies

		Substrate: Duroid 5880, $\epsilon_r = 2.2$, $h = 0.508$ mm.			
		Desired $Z_0 = 50$ Ω.			
f (GHz)	W (mm)	ϵ_{reff}	ΔL (mm)	L (mm)	W_0 (mm)
10	11.859	2.0876	0.267	9.848	1.565
20	5.929	2.0213	0.264	4.748	1.565

Table 2: Patch dimensions

(a) Structure Parameters

(b) S_{11} vs. frequency

(c) Smith chart

(d) Radiation pattern

Figure 4: Simulation results for 10 GHz patch antenna

(a) Structure Parameters

(b) S_{11} vs. frequency

(c) Smith chart

(d) Radiation pattern

Figure 5: Simulation results for 20 GHz patch antenna

	10 GHz antenna	20 GHz antenna
Directivity	7.66	7.91
Gain	7.30	7.43
Efficiency	91.89 %	89.53%
Maximum effective area	3.8462 cm^2	0.9907 cm^2
Bandwidth for $S_{11} = -10$ dB	208 MHz	1090 MHz

Table 3: Simulated antenna parameters.

Figure 6: Impedance matching design

is unknown, we design open circuit stubs which reflect the 20 GHz wave at the input and 10 GHz wave at the output. This creates a short circuit reference plane for 20 GHz at the input and for 10 GHz at the output. This type of reflector concept is used previously in FET frequency doublers [12].

In order to maximize the power transfer from the 10 GHz antenna to the 20 GHz antenna, we design input conjugate matching to maximize the power transfer of 10 GHz signal from the 10 GHz antenna to the diode. Output matching is also conjugate matching to maximize the power transfer of 20 GHz signal generated in the diode from the diode to the 20 GHz antenna. Double stub matching is used to provide adequate degrees of freedom to incorporate both impedance matching and reflector design. The requirements on the matching circuit are summarized below:

$$\Gamma_{L,10} = e^{j\theta_1}, \tag{14}$$
$$\Gamma_{S,20} = e^{j\theta_2}, \tag{15}$$
$$\Gamma_{in,10} = \Gamma_{S,10}^\star, \tag{16}$$
$$\Gamma_{out,20} = \Gamma_{L,20}^\star. \tag{17}$$

The subscripts 10 and 20 stand for 10 GHz and 20 GHz respectively.

Let us denote the diode S parameters at 10 GHz by $\{S_{11}, S_{12}, S_{13}, S_{14}\}$ and those at

20 GHz by $\{S'_{11}, S'_{12}, S'_{13}, S'_{14}\}$. We note that, because of the nonlinearity, the S-parameters depend on the signal level. The operating condition of the diode is assumed to have a power level of around -10 dBm. Using the diode model for the Agilent HSCH-9161 diode, the S-parameters were generated at both 10 GHz and 20 GHz for the matching network design. For conjugate matching,

$$\Gamma_{in,10} = \Gamma^\star_{S,10} = S_{11} + \frac{S_{11}S_{21}\Gamma_{L,10}}{1 - S_{22}\Gamma_{L,10}} \tag{18}$$

$$= S_{11} + \frac{S_{11}S_{21}e^{j\theta_1}}{1 - S_{22}e^{j\theta_1}} \tag{19}$$

$$\Gamma_{out,20} = \Gamma^\star_{L,20} = S'_{22} + \frac{S'_{12}S'_{21}\Gamma_{S,20}}{1 - S'_{11}\Gamma_{S,20}} \tag{20}$$

$$= S'_{22} + \frac{S'_{12}S'_{21}e^{j\theta_2}}{1 - S'_{11}e^{j\theta_2}} \tag{21}$$

The double stub design for output and input matching are shown in Fig. 7. For output matching (Fig. 7(a)), segment (1) has a quarter wavelength at 10 GHz and acts as a 10 GHz reflector. The rest of the segments are used to obtain matching for maximum power transfer at 20 GHz. Similarly, for input matching (Fig. 7(b)), segment (1) has a quarter wave length at 20 Ghz and acts as a reflector. The rest of the stubs are used to design the impedance matching for maximum power transfer at 10 GHz.

5   Measurement results

To measure the performance of the passive frequency doubler, the 10 GHz and 20 GHz patch antennas and several test structures were fabricated and tested. All the measurements were performed in an anechoic chamber.

Antenna	S_{11} (dB)	Gain (dB)
10 GHz	-15	4.8
20 GHz	-7	1.4

Table 4: Measured antenna parameters

The antenna measurements are shown in Fig. 8. Measured antenna gain and peak S_{11} are summarized in Table 4. The other structures which are fabricated are shown in Fig. 9. Fig. 9(a) shows the Agilent HSCH-9161 GaAs Schottky diode soldered in the 0.48 mm gap between two transmission lines. Another test structure is fabricated which includes the double stub impedance matching sections (Fig. 9(b)). This is fabricated to access the effect of matching. These circuits are provided with an SMA connector to be able to measure

(a) Output impedance match

(b) Input impedance match

Figure 7: Impedance matching and reflector design

(a) 10 GHz antenna

(b) S_{11}

(c) Radiation Pattern

(d) 20 GHz antenna

(e) S_{11}

(f) Radiation Pattern

Figure 8: Antenna measurements

(a)

(b)

(c)

Figure 9: (a) Transmission line test structure with a Schottky diode connected between a gap in the transmission lines (b) Diode with double stub impedance matching sections (c) Complete passive doubler with integrated antennas.

the losses in the circuit for a given input power. Fig. 9(c) shows the photo of the complete integrated frequency doubler.

Fig. 10 shows the measurement results of 20 GHz signal power for different for various power levels of the input 10 GHz signal for the test structures in Fig. 9(a) - 9(b). It shows that the device with double stub matching has better second harmonic than the design without impedance matching. For example, at a power level of −10 dBm of the input 10 GHz signal, the power generated at 20 GHz is −30.0 dBm for the structure in Fig. 9(b) and −45.6 dBm for the structure in Fig. 9(a). This means that double stub impedance matching design can give another 15.6 dB gain than no matching design when the input power level is -10.0 dBm. We note that the overall conversion loss is significant. The plot also illustrates the increasing effect of nonlinearity as the input power level is increased.

Figure 10: Performance comparison between structures in Fig. 9(a) and 9(b).

For the structure in Fig. 9(c), a 10 GHz horn antenna (Rozendal Associates P/N RA-4610-6 S/N 004) with a 15 dB gain is used to illuminate the harmonic reradiator. The reradiated power at 20 GHz is received using another horn antenna (Microwave Associates standard horn model 653-6) having a 14 dB gain. From the measurements of the reradiated power levels, the scattering cross-section of the structure is calculated. The scattering cross-section is defined such that the effective isotropic radiated power at the harmonic is the product of the incident power density at the fundamental and the scattering cross-section. This is shown in Table 5. The effective isotropic power (EIRP) radiated by the harmonic reradiator is also shown. The ideal cross-section in the table represents the cross-section presented by the harmonic reradiator if all the available power at the fundamental is reradiated at the harmonic (for the given antenna configuration). Owing to the limitations imposed by the dynamic range of the measurement equipment, measurements are taken over a limited range of power densities. An example energy budget is presented in the next section to show that the power densities are practical.

Ideal cross section	2.95 cm^2	
Power density (S) (W/m^2)	EIRP (dBm)	$\sigma(S)$ (cm^2)
0.10	-35.2	2.98×10^{-2}
0.16	-33.6	2.72×10^{-2}
0.25	-31.2	2.98×10^{-2}
0.40	-29.2	2.99×10^{-2}
0.80	-25.6	3.43×10^{-2}

Table 5: Harmonic reradiator cross-section and EIRP

6   Example energy budget

Consider a radar with a peak transmit power of 50 KW. Let the gain of the antenna be 36 dB. For example, a uniformly illuminated circular aperture of 0.6 m diameter has such a gain. Then the power density at a distance of 10 Km is approximately 0.158 W/m^2. From the Table 5, we see that the cross-section presented by the sensor at the second harmonic is 2.98×10^{-2} cm^2. The same radar antenna has a gain of 42 dB at the second harmonic. The power received by the radar is -130 dBm. To detect the presence of a sensor the radar compares the energy received during the observation period to a threshold. The performance depends on the ratio of the signal energy to the noise-variance. Assuming that the noise at the receiver is white and the noise temperature is 1000 K (representing a noisy receiver), to obtain a post processing signal to noise ratio (Ratio of the energy received and the input noise spectral density) of 10 dB, the radar should transmit at its peak power for a duration of 2 ms. Radars can typically illuminate the sensor for a longer duration. This demonstrates that the designed frequency doubler can indeed be a valuable communicating node in remote sensing applications.

7   Conclusion

We have successfully designed a passive second harmonic reradiator capable of reradiating harmonics at low incident power densities. The design utilizes the nonlinearity of a diode and does not require external power supply. The harmonic reradiator we demonstrate should be applicable in a remote sensing network, and could serve as a basis for a small and low cost communication device.

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