
This is the **accepted version** of the journal article:

Casacuberta Orta, Pau; Vélez Rasero, Paris; Su, Lijuan; [et al.]. «Highly sensitive lossy tunable permittivity sensor». IEEE Microwave and Wireless Technology Letters, Vol. 34, issue 6 (June 2024), p. 687-690. DOI 10.1109/LMWT.2024.3383162

This version is available at <https://ddd.uab.cat/record/294899>

under the terms of the  ^{IN}
COPYRIGHT license

Highly Sensitive Lossy Tunable Permittivity Sensor

Pau Casacuberta, *Graduate Student Member, IEEE*, Paris Véléz, *Senior Member, IEEE*, Lijuan Su, *Member, IEEE*, Xavier Canalias, and Ferran Martín, *Fellow, IEEE*

Abstract—This paper presents a novel lossy tunable permittivity sensor that enhances sensitivity through a controllable conductivity method using a junction field-effect transistor (JFET) in the ohmic region. The sensor, a one-port reflective-mode device, includes a coplanar waveguide (CPW) ending in a step impedance resonator (SIR) as the sensing element, with the JFET bridging the SIR patch and ground. The high sensitivity is achieved by virtue of the strong shift in the phase of the reflection coefficient at resonance experienced by varying the dielectric constant of the material under test (MUT). This device not only serves as a permittivity sensor but also for tracking variables linked to permittivity changes. Our experimental results uniquely demonstrate that sensitivity to minute perturbations can be significantly enhanced by tuning losses. The prototype, designed and constructed for detecting submillimeter proximity changes with a dielectric slab, achieves a peak sensitivity of $2291^\circ/\text{mm}$ within a compact $8.5 \text{ mm} \times 12 \text{ mm}$ footprint.

Index Terms—coplanar waveguide (CPW), JFET, lossy sensor, permittivity sensor, phase-variation sensor, tunable sensor.

I. INTRODUCTION

Achieving ultra-high sensitivity in planar microwave sensors with compact sensing regions presents a significant challenge. These sensors, which primarily detect changes in the dielectric properties of their surroundings, are versatile due to these properties' association with various physical, chemical, and biological variables. Consequently, microwave permittivity sensors find applications across a broad spectrum, including but not limited to temperature and humidity monitoring, concentration measurements of electrolytes and glucose, and detection of food deterioration and bacterial growth [1]–[12].

The performance of planar microwave sensors is governed by their operating principle, most commonly frequency variation. In this approach, changes in the measurand cause shifts in the resonance frequency, magnitude, or quality factor of the sensor. While these sensors offer robustness to electromagnetic interference (EMI) and noise, their sensitivity is inherently constrained unless operated at very high frequencies. Additionally, they require broadband input signals to achieve the necessary output dynamic range [13].

Recent advancements in planar microwave sensors, particularly one-port reflective-mode phase-variation sensors

[12], [14]–[18], have demonstrated unprecedented sensitivities. These sensors operate by measuring the phase of the reflection coefficient at a specific frequency, and provide high sensitivities as far as a high phase slope is achieved at that frequency [19]. This goal has been realized through the use of either distributed or semi-lumped resonators, including high-impedance quarter-wavelength and low-impedance half-wavelength open-ended resonators [14], step-impedance resonators with high inductance-to-capacitance ratios [15], or slot resonators with low inductance-to-capacitance ratios [16]. Enhancing the phase slope, and thereby the sensitivity, can also be achieved by cascading impedance inverters with alternating high and low characteristic impedance between the input port and the sensing element [14]. Alternatively, high sensitivity can be reached through the use of a pair of weakly coupled resonators, either distributed or semi-lumped, which leads to a large phase excursion between two closely spaced resonance frequencies, with the operational frequency finely tuned in between [17], [18].

II. THE PROPOSED LOSSY SENSOR AND ANALYSIS

The rationale behind the need of a high phase slope (at the operating frequency, f_0) to achieve high sensitivity is very simple: when the permittivity of the immediate environment surrounding the sensor varies, the resulting downward or upward shift in the overall phase response generates a phase excursion at f_0 that increases with the phase slope. Thus, a high phase slope boosts the sensitivity, or the derivative of the phase of the reflection coefficient with respect to the input variable (typically, the dielectric constant of the so-called material under test (MUT) or a variable related to it). Let us next obtain the phase slope of a lossy resonator with impedance $Z_L(\omega) = R(\omega) + j\chi(\omega)$, where $R(\omega)$ and $\chi(\omega)$ are the real and the imaginary part, respectively, and ω is the angular frequency. The reflection coefficient is

$$\rho = \frac{Z_L - Z_0}{Z_L + Z_0} \quad (1)$$

and hence the phase of the reflection coefficient can be expressed as

$$\phi_\rho = \arctan\left(\frac{\chi(\omega)}{R(\omega) - Z_0}\right) - \arctan\left(\frac{\chi(\omega)}{R(\omega) + Z_0}\right). \quad (2)$$

The phase slope evaluated at resonance, where $\chi(\omega) = 0$, or $\omega = \omega_0 = 2\pi f_0$, is found to be

$$\left.\frac{d\phi_\rho}{d\omega}\right|_{\omega=\omega_0} = \frac{2Z_0}{R^2(\omega_0) - Z_0^2} \cdot \left.\frac{d\chi(\omega)}{d\omega}\right|_{\omega=\omega_0}. \quad (3)$$

From (3), approaching the matching condition ($R(\omega_0) \rightarrow Z_0$), the phase slope at resonance approaches

Manuscript received April 19, 2021; revised August 16, 2021.

This work was supported by MCIN/AEI 10.13039/501100011033, Spain, projects PID2019-103904RB-I00, PID2022-139181OB-I00 (ERDF European Union), and PDC2021-121085-I00 (European Union Next Generation EU/PRTR), by AGAUR, Catalonia Government, project 2021SGR00192, and by ICREA. P. Casacuberta and X. Canalias acknowledge the Ministerio de Universidades, Spain, for the grants FPU20/05700 and PREP2022-000364, respectively. L. Su acknowledges the Juan de la Cierva Program for the support through the project IJC2019-040786-I. P. Véléz acknowledges the Ramon y Cajal Program for the support through the project RYC2022-035819-I.

The authors are with CIMITEC, Departament d'Enginyeria Electrònica, Universitat Autònoma de Barcelona, Spain (e-mail: pau.casacuberta@uab.cat; ferran.martin@uab.cat).

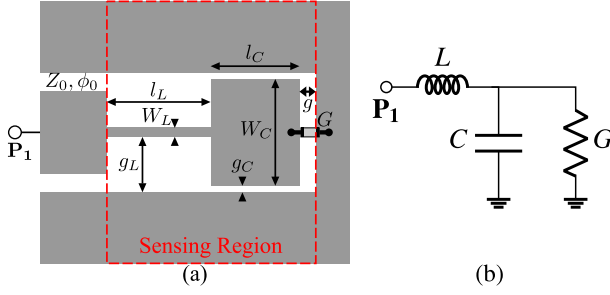


Fig. 1. Sensor topology (a) and circuit model (b). Relevant dimensions in mm (for the prototype device) are: $W_L = 0.3$, $l_L = 5$, $g_L = 1.85$, $W_C = 3.6$, $l_C = 3$, $g_C = 0.2$, $g = 0.5$.

infinity, enhancing sensitivity as losses are engineered. However, perfect matching ($[R(\omega_0) = Z_0]$) is to be avoided to prevent complete power absorption at f_0 by the resonant element. This new sensitivity optimization method, focusing on engineered losses, eliminates the need for cascading impedance inverters or adding a weakly coupled resonator.

Fig. 1 shows the lossy sensor's typical topology and its equivalent circuit model, featuring a coplanar waveguide (CPW) ending in a step impedance resonator (SIR) with a tunable resistor (with conductance G_L). The SIR's narrow strip inductance is represented by L , and its patch capacitance by C . Losses from the substrate and the material under test (MUT) are included in G_S . The impedance seen from the input port (by excluding the access line) is

$$Z_L = R(\omega) + j\chi(\omega) = \frac{G}{G^2 + \omega^2 C^2} + j\omega \left(L - \frac{C}{G^2 + \omega^2 C^2} \right). \quad (4)$$

Introducing $R(\omega)$ and $\chi(\omega)$ in (3), and evaluating at the resonance frequency, ω_0 , provides

$$\left. \frac{d\phi_p}{d\omega} \right|_{\omega_0} = \frac{4Z_0 L \left(1 - \frac{G^2 L}{C} \right)}{\frac{G^2 L^2}{C^2} - Z_0^2} \equiv P \quad (5)$$

where $G = G_S + G_L$ and the resonance frequency is

$$\omega_0 = \sqrt{\frac{1}{LC} - \frac{G^2}{C^2}}. \quad (6)$$

According to (5), there is a pole in the phase slope at resonance, P . Such pole corresponds to a value of G given by

$$G_p = Z_0 \frac{C}{L}. \quad (7)$$

For $L/C > Z_0^2$, the typical situation to enhance the sensitivity [15], the pole is lower than the zero, located at

$$G_z = \sqrt{\frac{C}{L}}. \quad (8)$$

Thus, P is negative before the pole, and it takes the value

$$P = -\frac{4L}{Z_0} \quad (9)$$

when losses are negligible ($G = 0$). The magnitude of P increases as G nears the pole. High L (thus, low C) boosts the phase slope. Setting G close to the pole enhances sensitivity.

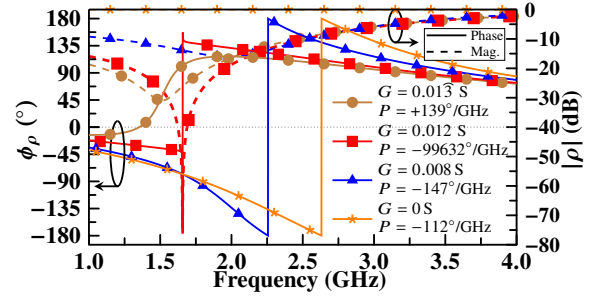


Fig. 2. Phase and magnitude response inferred by circuit simulation for the structure of Fig. 1 for four cases: (i) $G = 0 \ll G_p$, (ii) $G = 0.008 \text{ S} \leq G_p$, (iii) $G = 0.012 \text{ S} = G_p$, and (iv) $G = 0.013 \text{ S} \geq G_p$. The port's reference impedance is $Z_0 = 50\Omega$. At resonance, the phase is either $\phi_p = \pm 180^\circ$ (for $G < G_p$) or $\phi_p = 0^\circ$ (for $G > G_p$), and the phase slope P is indicated.

III. SENSOR DESIGN AND EXPERIMENTAL VALIDATION

The sensor was built on a *Rogers 4003C* substrate ($\epsilon_r = 3.55$, thickness $h = 1.524\text{mm}$, $\tan \delta = 0.0022$), dimensions detailed in Fig. 1's caption. With these specifications, the sensor's resonance frequency is $f_0 = 2.636 \text{ GHz}$, both without a resistor ($G_L = 0 \text{ S}$) and with the reference material ($\epsilon_{\text{REF}} = 3.55$), as confirmed by electromagnetic simulations. The sensor's reactive parameters, $L = 3.89 \text{ nH}$ and $C = 0.94 \text{ pF}$, were derived assuming negligible G_S , following the approach in [15]. When losses ($G_L \neq 0$) are considered, part of the signal is absorbed by the resonator, impacting the reflection coefficient at resonance, as shown by

$$\rho|_{\omega_0} = \frac{\frac{G}{G^2 + \omega_0^2 C^2} - Z_0}{\frac{G}{G^2 + \omega_0^2 C^2} + Z_0} = \frac{\frac{GL}{C} - Z_0}{\frac{GL}{C} + Z_0} \quad (10)$$

and the magnitude of it decreases as G gets closer to the pole.

Figure 2 shows reflection coefficient of Fig. 1(b) by means of *Keysight ADS* circuitual simulation, using the specified reactive parameters and varying conductance values G , affecting phase slopes at frequency f_0 . With a reference impedance $Z_0 = 50\Omega$, the critical conductance $G_p = 0.012 \text{ S}$ marks the transition between different regimes. Below G_p , the system is in a low-loss state with a negative phase slope at f_0 . At G_p , the phase slope becomes infinite, and above G_p , it enters a high-loss state with a positive phase slope. Approaching G_p enhances the phase slope at resonance, decreases the reflection coefficient's magnitude, and signifies a phase slope sign change beyond G_p . These effects underscore the role of engineered losses in optimizing phase slope and sensitivity by adjusting G (it is convenient to set G slightly below G_p).

To validate our sensitivity enhancement approach, we fabricated the device depicted in Fig. 1 using an *LPKF H100* milling machine and performed two tests. Initially, resistors with different conductance values were inserted between the capacitive patch and ground plane, as shown in Fig. 1, soldered behind the substrate to connect with the top metal layers by means of vias, avoiding MUT interaction. We created four distinct devices, each featuring a specific resistor value (or G), showcased in Fig. 3. The reflection coefficients' phases and magnitudes were measured for these devices with a *Keysight 5221A* vector network analyzer (VNA), results shown in Fig. 4.

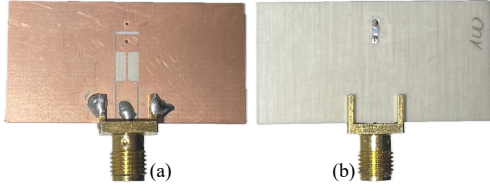


Fig. 3. Photograph of one of the fabricated sensors with a resistor connected between the SIR patch and the ground plane. (a) Top view; (b) bottom view.

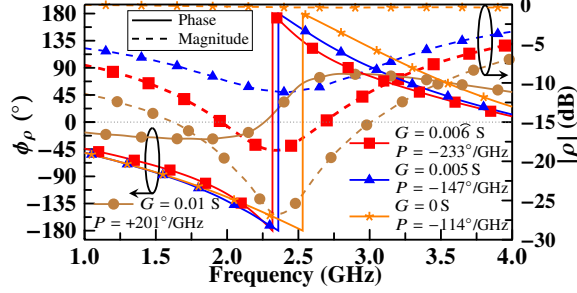


Fig. 4. Measured phase and magnitude of the reflection coefficient for the sensor of Fig. 3 with a set of resistances, available in our lab, corresponding to the indicated conductance values, including the resonance phase slope, P .

Despite vias and resistor imperfections affecting the theoretical model prediction, these results confirm conductance's role in boosting the sensitivity to smooth perturbations, a topic we expand on later.

Utilizing the four sensors depicted in Fig. 3, submillimeter-scale proximity measurements were conducted with a vertically maneuverable dielectric slab above the SIR, affecting the effective dielectric constant and thus the phase of the reflection coefficient. The baseline or REF position, with the MUT (a stack of microwave substrates of dielectric constant $\epsilon_{\text{SLAB}} = \epsilon_{\text{REF}} = 3.55$ and thickness $h = 10$ mm) atop the SIR, was used to align the operating frequency with the SIR's resonance at a specific conductance G , detailed in Fig. 4. The reflection phase at resonance measurements showed that sensitivity to slab's vertical distance increases as conductance nears G_p , illustrated in Fig. 5. This confirms that modifying losses can notably improve sensitivity to minor displacements.

The second validation featured a voltage-tunable JFET (SKY65050-372LF) connected between the SIR patch and ground, allowing conductance adjustment via gate voltage (V_G) for tunable loss-assisted sensitivity. Fig. 6 shows the sensor and its circuit model with the JFET. VNA measurements with different V_G values (Fig. 7) demonstrate tunable sensitivity and phase slope at resonance. Phase variation measurements of the reflection coefficient with slab distance, shown in Fig. 5, confirm the method's ability to significantly boost sensitivity up to 2291 $^\circ/\text{mm}$ in the limit of small displacements at $V_G = -697$ mV. This technique maintains the sensor's small footprint (8.5 mm $\times$ 12 mm) without needing impedance inverters for sensitivity enhancement.

IV. CONCLUSION

This research introduces a permittivity sensor with significant detection capabilities, achieving 2291 $^\circ/\text{mm}$, far surpass-

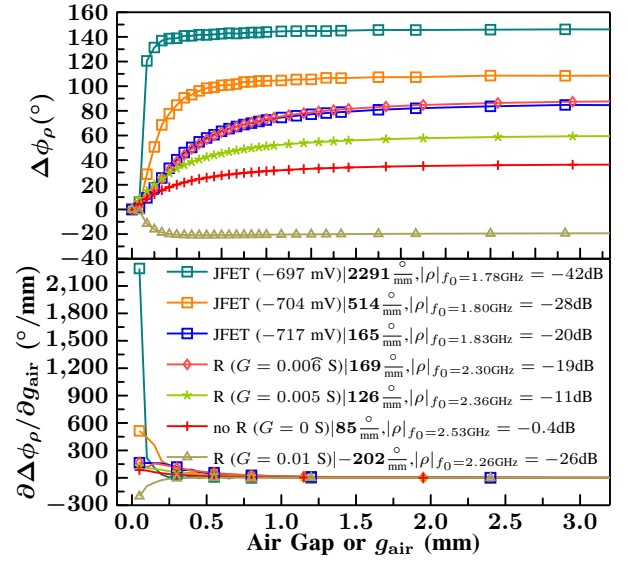


Fig. 5. Measured dependence of the phase of the reflection coefficient at resonance with the vertical distance, g_{air} , and sensitivity, for the sensors of Fig. 3 and 6 for the indicated cases. For vertically displacing the slab, the Thorlabs MVS010/M displacement system was used. The indicated sensitivities correspond to the limit of small air gaps (or $g_{\text{air}} = 0$ mm).

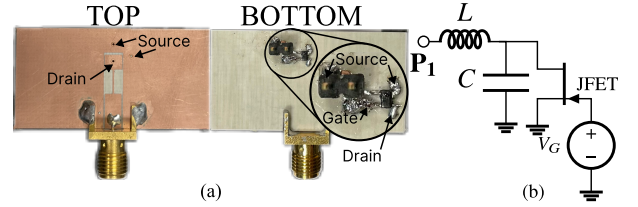


Fig. 6. Photograph (a) and circuit model (b) of the JFET lossy tunable sensor.

ing the benchmarks of 110 $^\circ/\text{mm}$ and 70 $^\circ/\text{mm}$ reported in [14] and [17], respectively. The cornerstone of this breakthrough is the sensor's tunability, achieved through a voltage-controlled JFET operating in the ohmic region. This allows for unparalleled flexibility in sensitivity adjustment, accommodating a wide array of scenarios, without necessitating individual redesigns for different applications. Future investigations will delve into a detailed sensitivity analysis incorporating nonidealities to broaden the scope of this innovative approach.

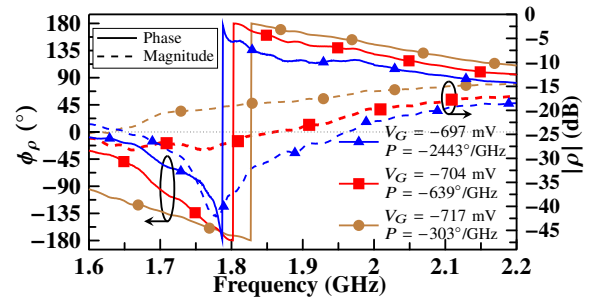


Fig. 7. Measured phase and magnitude of the reflection coefficient for the circuit of Fig. 6(b) by considering different bias voltages. Non-idealities are due to the biasing circuit surrounding the JFET and the transistor itself.

REFERENCES

- [1] F. Martín, P. Vélez, J. Muñoz-Enano, and L. Su, *Planar Microwave Sensors*. Wiley, Sep. 2022.
- [2] D. Girbau, A. Ramos, A. Lazaro, S. Rima, and R. Villarino, “Passive wireless temperature sensor based on time-coded UWB chipless RFID tags,” *IEEE Trans. Microw. Theory Tech.*, vol. 60, no. 11, pp. 3623–3632, Nov. 2012.
- [3] W. M. Abdulkawi and A.-F. A. Sheta, “Chipless RFID sensors based on multistate coupled line resonators,” *Sens. Actuators Phys.*, vol. 309, p. 112025, Jul. 2020.
- [4] J. Yeo, J.-I. Lee, and Y. Kwon, “Humidity-sensing chipless RFID tag with enhanced sensitivity using an interdigital capacitor structure,” *Sensors*, vol. 21, no. 19, p. 6550, Sep. 2021.
- [5] E. M. Amin, M. S. Bhuiyan, N. C. Karmakar, and B. Winther-Jensen, “Development of a low cost printable chipless RFID humidity sensor,” *IEEE Sens. J.*, vol. 14, no. 1, pp. 140–149, Jan. 2014.
- [6] J. Muñoz-Enano *et al.*, “Characterization of electrolyte content in urine samples through a differential microfluidic sensor based on dumbbell-shaped defected ground structures,” *Int. J. Microw. Wirel. Technol.*, vol. 12, no. 9, pp. 817–824, Apr. 2020.
- [7] C. G. Juan *et al.*, “Concentration measurement of microliter-volume Water–Glucose solutions using Q factor of microwave sensors,” *IEEE Trans. Instrum. Meas.*, vol. 68, no. 7, pp. 2621–2634, Jul. 2019.
- [8] —, “Study of q_u -based resonant microwave sensors and design of 3-d-printed devices dedicated to glucose monitoring,” *IEEE Transactions on Instrumentation and Measurement*, vol. 70, pp. 1–16, 2021.
- [9] J. Muñoz-Enano, P. Vélez, P. Casacuberta, L. Su, and F. Martín, “Characterization of edible oils subjected to industrial frying processes through high sensitivity microwave sensors,” in *2023 53rd Eur. Microw. Conf. EuMC*. IEEE, Sep. 2023.
- [10] R. Narang *et al.*, “Sensitive, real-time and non-intrusive detection of concentration and growth of pathogenic bacteria using microfluidic-microwave ring resonator biosensor,” *Sci. Rep.*, vol. 8, no. 1, Oct. 2018.
- [11] M. C. Jain, A. V. Nadaraja, R. Narang, and M. H. Zarifi, “Rapid and real-time monitoring of bacterial growth against antibiotics in solid growth medium using a contactless planar microwave resonator sensor,” *Sci. Rep.*, vol. 11, no. 1, Jul. 2021.
- [12] J. Munoz-Enano, P. Velez, L. Su, M. Gil-Barba, and F. Martin, “A reflective-mode phase-variation displacement sensor,” *IEEE Access*, vol. 8, pp. 189 565–189 575, 2020.
- [13] C.-L. Yang, C.-S. Lee, K.-W. Chen, and K.-Z. Chen, “Noncontact measurement of complex permittivity and thickness by using planar resonators,” *IEEE Trans. Microw. Theory Tech.*, vol. 64, no. 1, pp. 247–257, Jan. 2016.
- [14] J. Munoz-Enano *et al.*, “On the sensitivity of reflective-mode phase-variation sensors based on open-ended stepped-impedance transmission lines: Theoretical analysis and experimental validation,” *IEEE Trans. Microw. Theory Tech.*, vol. 69, no. 1, pp. 308–324, Jan. 2021.
- [15] P. Casacuberta *et al.*, “Circuit analysis of a coplanar waveguide (CPW) terminated with a step-impedance resonator (SIR) for highly sensitive one-port permittivity sensing,” *IEEE Access*, vol. 10, pp. 62 597–62 612, 2022.
- [16] J. Muñoz-Enano, P. Vélez, P. Casacuberta, L. Su, and F. Martín, “Reflective-mode phase-variation permittivity sensor based on a step-impedance microstrip line terminated with a slot resonator for solid and liquid characterization,” *IEEE Trans. Microw. Theory Tech.*, pp. 1–0, 2023.
- [17] P. Casacuberta, P. Vélez, J. Muñoz-Enano, L. Su, and F. Martín, “Highly sensitive reflective-mode phase-variation permittivity sensors using coupled line sections,” *IEEE Trans. Microw. Theory Tech.*, vol. 71, no. 7, pp. 2970–2984, Jul. 2023.
- [18] —, “Highly sensitive coplanar waveguide (CPW) reflective-mode phase-variation permittivity sensors based on weakly coupled step-impedance resonators (SIRs),” *IEEE Trans. Microw. Theory Tech.*, pp. 1–15, 2023.
- [19] —, “Losses assisted sensitivity enhancement in reflective-mode phase-variation permittivity sensors based on weakly coupled distributed resonators,” *IEEE Sens. Lett.*, vol. 7, no. 8, pp. 1–4, Aug. 2023.