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Low Cost *X*-Band MIC BARITT Doppler Sensor

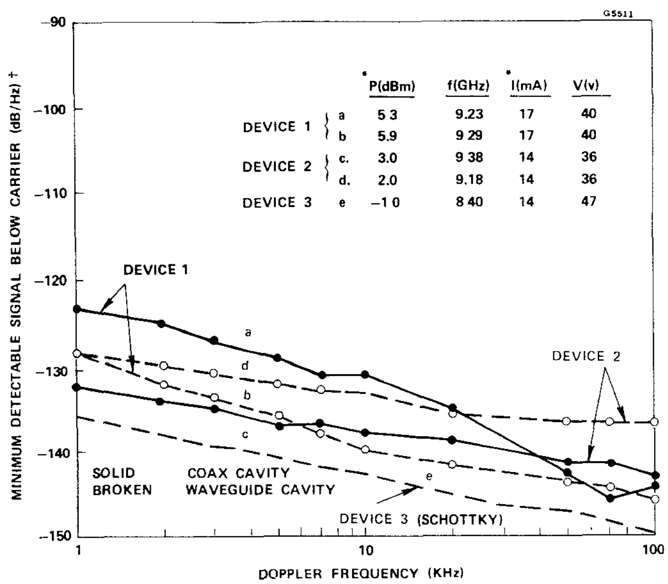
SIANG PING KWOK, MEMBER, IEEE, AND KENNETH P. WELLER, MEMBER, IEEE

***Abstract*—An MIC *X*-band BARITT self-mixing oscillator has been developed. A minimum detectable signal below carrier of −139 dB/Hz at 100 kHz away from carrier was achieved at 1-mW signal carrier. A low cost, compact and sensitive hybrid MIC Doppler sensor module was constructed, incorporating the BARITT MIC circuit and a microstrip antenna.**

I. Introduction

THE superior self-mixing property of the BARITT device offers potential advantages for Doppler sensor applications over IMPATT and Gunn devices [1]–[3]. However, a major factor inhibiting the widespread use of microwave Doppler sensors is the cost of reliable microwave components. Microwave integrated circuit (MIC) technology using microstrip offers advantages for miniaturization, reliability, and low cost.

This paper describes the design of the BARITT device, the MIC oscillator circuit, the sensitivities of the self-oscillating mixer, and a hybrid MIC *X*-band Doppler sensor module, incorporating the BARITT MIC circuit and a microstrip antenna.



*P, I for maximum MDSBC.
† for S+N/N = 3 dB.

Fig. 1. BARITT self-mixing sensitivity.

II. Device Design and Self-Mixing Sensitivity

There are two figures of merit that can be applied to self-oscillating detectors. One is the minimum detectable signal (MDS) level which is related directly to the noise figure of the detector. For a self-mixing oscillator, the output power may be relatively low for the best MDS. The second figure of merit is the minimum detectable signal below carrier (MDSBC). It is directly related to the maximum range capability of the system, since it is the ratio of transmitted power to minimum detectable received signal. Experimentally, it was found that an excessively large oscillator output power may actually reduce the MDSBC ratio, as shown in Fig. 1. It is suggested that the oscillator noise and conversion loss are the dominant factors. Thus the BARITT device design for this application was aimed primarily at maximizing the oscillator MDSBC and not its power output.

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S. P. Kwok is with the Torrance Research Center, Hughes Aircraft Company, Torrance, CA 90509.

K. P. Weller was with the Torrance Research Center, Hughes Aircraft Company, Torrance, CA. He is now with TRW Inc., Redondo Beach, CA 90278.

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Because the silicon p^+-n-p^+ BARITT structure exhibits a higher negative conductance than its n^+-p-n^+ counterpart [2], [4], the former is preferred for ease of operation in the MIC circuit. Based on an operating frequency of 10 GHz and a desired operating voltage under 50 V, the doping density and thickness of the n-type epitaxial base region were chosen to be 1.5×10^{15} cm^{-3} and 5.2 μm, respectively.

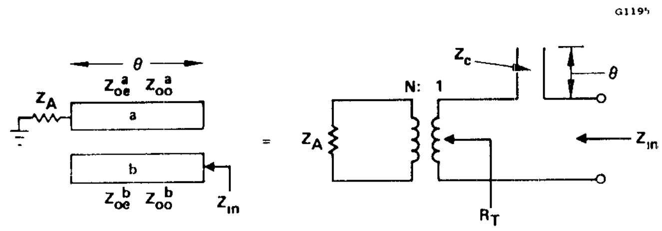
Fig. 1 shows the self-mixing sensitivity (MDSBC) of two BARITT devices operated in a coaxial and waveguide cavity. The injecting junction of Device 1 was formed by a shallow boron diffusion thus forming an abrupt p^+-n junction. Device 2 was subjected to additional diffusion drive cycle resulting in a more graded injecting junction. As apparent from Fig. 1, Device 2 exhibits less $1/f$ noise than Device 1. The Schottky-barrier junction (Device 3) exhibits the best self-mixing sensitivity, -135 dB/Hz at 1 kHz and -150 dB/Hz at 100 kHz. It is evident that the device with a Schottky barrier, though operated at the minimum power (-1 dBm) compared to the rest of the devices, exhibits the best MDSBC ratio.

III. BARITT MIC Circuit Design

The inherent problem with the BARITT device is that its series equivalent negative resistance is very small, typically less than 1 Ω, one order of magnitude smaller than those of the Gunn and IMPATT devices. The impedance of the BARITT device essentially consists of a depletion capacitance susceptance B shunted with a small negative conductance ($-B/Q$). According to large-signal numerical calculations the typical large-signal Q of an X-band p^+-n-p^+ structure is in the order of -100 [1], [2]. The structure used here has a base width of approximately 5 μm and cross-sectional diameter of 1.8×10^{-2} cm. The corresponding device series equivalent impedance at 10 GHz is $-0.3-j30$ Ω. This value was in agreement with the measurements obtained from coaxial circuits under optimum power and efficiency conditions. In order to match the diode chip impedance to a 50-Ω load, two impedance transformation steps were taken. The diode package lead inductance was first used to resonate with the diode capacitance and an impedance transformation circuit was then employed to match the small real-part diode impedance to the 50-Ω load.

An MIC coupled-line circuit configuration was chosen for two reasons. 1) The circuit is capable of transforming relatively high load impedance (50 Ω) to a very small value of 1 Ω or less over a wide-band frequency. 2) The bias circuit is isolated from the RF output. The basic circuit configuration and its equivalent circuit is depicted in Fig. 2. The diode is placed at the Z_{in} end of line b. The 50-Ω output load is at Z_A. The load impedance is transformed to a lower value R_T by a ratio of N^2, shunted with an open stub of an electrical length equal to that of the parallel strips [5], [6].

For Duroid ($\epsilon_r=2.34$) substrate the coupled linewidth-to-height-ratio (W/H) and spacing-to-height ratio (S/H) have been calculated as a function of even-mode characteristic impedance Z_{oe} with a constraint that $Z_{oo}+Z_{oe}=2Z_A$, where Z_{oo} is the odd-mode characteristic impedance. (W/H) and (S/H) and this corresponding transformed impedance R_T are plotted in Fig. 3. To transform a 50-Ω load to 1 Ω or less, the value of Z_{oe} lies in the range of 52–58 Ω. Minor modifications to this basic design were made empirically for optimum-device operation. The final circuit pattern is shown in Fig. 4. The gap spacing S was reduced while the linewidth W was increased. This results in a tighter coupling to the output and reduced radiation loss. The diode is positioned a short distance l_2 from the coupled-line transformers to provide the proper inductive reactance in series with the load resistance. An approximate quarter-wavelength shunt open stub was introduced to the right of the diode to compensate for the increased circuit impedance presented to the diode as a result of the tighter coupling mentioned above. A Chebyshev low-pass



$$Z_{oe}^a = Z_A\left[\left(\frac{R_T}{Z_A}\right)^{\frac{1}{2}}+1\right]$$

$$Z_{oo}^a = 2Z_A - Z_{oe}^a$$

$$Z_{oo}^b = Z_S - Z_A\left[1-\frac{R_T}{Z_A}\right] + Z_{oo}^a$$

$$Z_{oe}^b = Z_{oo}^b + Z_{oe}^a - Z_{oo}^a$$

$$Z_c = Z_A\left[\frac{N^2-1}{N^2}\right] + Z_{oo}^b - Z_{oo}^a$$

$$N = \text{Turns Ratio} = \frac{2Z_A}{Z_{oe}^a - Z_{oc}^a}$$

$$R_T = \frac{Z_A}{N^2}$$

Fig. 2. Coupled-line circuit and its equivalent.

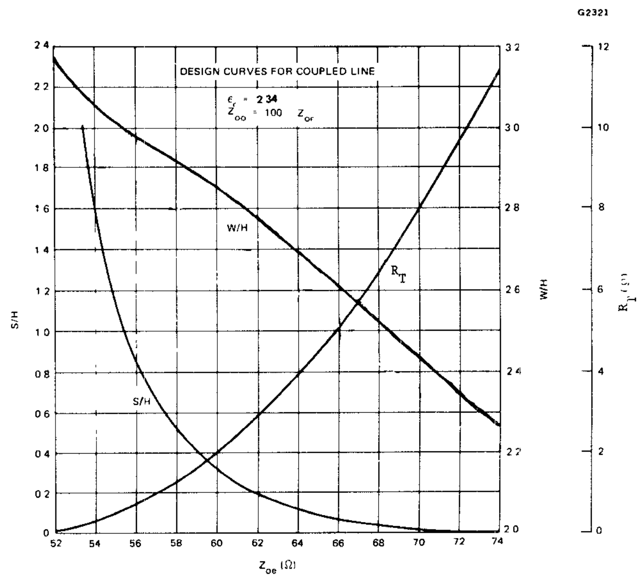


Fig. 3. Design curves for coupled-line circuit.

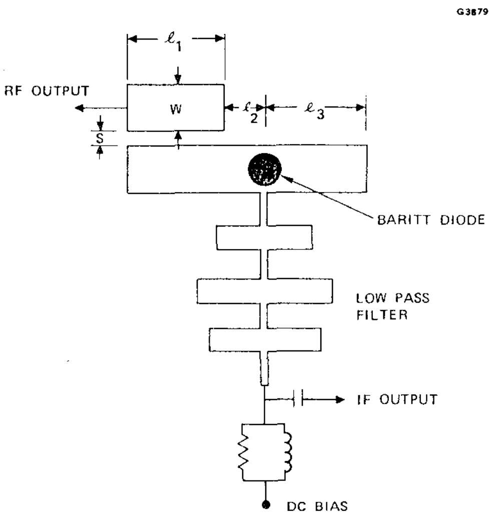


Fig. 4. Circuit layout for the MIC sensor module.

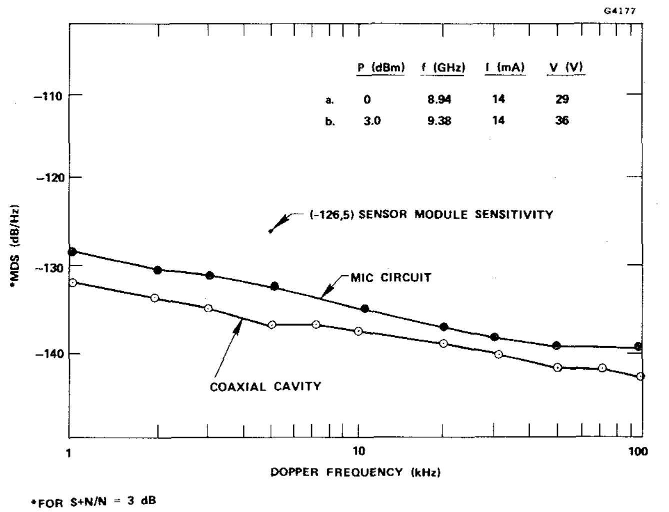


Fig. 6. Minimum detectable signal below carried as a function of Doppler frequency.

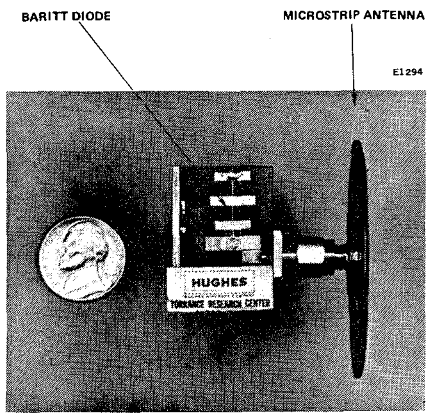


Fig. 5. BARITT MIC sensor circuit.

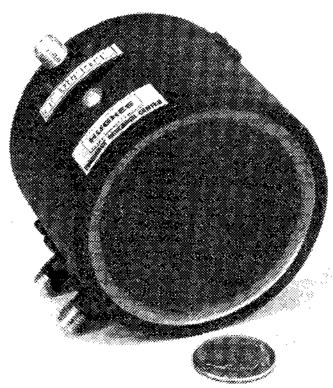


Fig. 7. BARITT Doppler sensor module.

filter is provided to prevent RF leakage into the bias line while permitting passage of the dc bias and IF signal generated by the self-mixing BARITT device.

A photograph of the completed MIC BARITT sensor is shown in Fig. 5. The Duroid substrate is bonded to a mounting block. The output is connected to a microstrip antenna through an OSM connector. The packaged BARITT device is held in a heat sink which screws into the mounting block from the bottom. The diode projects into a hole drilled through the Duroid substrate and is contacted on top by spring fingers soldered to the microstrip line. This fixed-tuned circuit provides 0.5–1.0 mW RF power at the OSM output connector with the low-power BARITT devices tested to date. This is on the order of 3 dB less power than can be obtained in the coaxial circuit, indicating that further optimization of the MIC circuit is possible. Radiation loss accounts for at least part of the decreased power in the open-microstrip circuit. The minimum detectable signal below carrier of the MIC BARITT oscillator is plotted as a function of Doppler frequency in Fig. 6. The resultant sensitivity is comparable to that obtained in the coaxial cavity. A MDSBC -139 dB/Hz (for system $(S+N)/N$ ratio of 3 dB) was achieved for Doppler frequencies beyond 100 kHz.

IV. Doppler Sensor Module

Fig. 7 shows the complete MIC BARITT Doppler sensor module. It consists of a microstrip planar antenna, an MIC BARITT self-mixing oscillator mounted on a heat-sink slab, a high-gain bandpass IF amplifier, and a threshold indicator. An envelope detector, a voltage comparator, and a light emitting diode (LED) form the threshold indicator circuitry.

The BARITT device and the IF circuitry are operated from -62.5 V and 12.5 V batteries, respectively. The output of BARITT oscillator was -1 dBm at X band. The BARITT device was biased at 22 mA and 32 V.

The antenna was fabricated on a 0.03-in$\times$2.00-in diameter Duroid microstrip. It consists of four microstrip dipole elements fed by a corporate feed network. Input to the microstrip is provided by an OSM flange mount connector. The MDSBC of -126-dB/Hz Doppler frequency was obtained in the final module.[1] This value reflects some degradation of BARITT self-mixing sensitivity from -132 dB/Hz achieved in the circuit shown in Fig. 6. This was partly due to nonoptimized device-circuit condition. The circuit had to be adjusted to match the resonant frequency of the narrow-band microstrip antenna.

V. Conclusion

An X-band MIC BARITT self-mixing oscillator has been developed with detection sensitivity comparable to that obtained with the coaxial cavity. A minimum detectable signal of -139 dB/Hz below carrier was achieved at 100 kHz away from the carrier. A compact, low cost and sensitive hybrid MIC Doppler sensor module has been constructed incorporating the BARITT MIC circuit and a microstrip antenna.

[1]The MDSBC was obtained from the measured range capability of the Doppler radar module.

Acknowledgment

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On the Use of a Microstrip Three-Line System as a Six-Port Reflectometer

RICHARD J. COLLIER AND NABIL A. EL-DEEB

***Abstract*—The scattering parameters for a coupled symmetrical three-line system in an inhomogeneous dielectric medium (e.g., microstrip) are derived directly in terms of a set of three orthogonal modes. The obtained results show that the condition for isolation of nonadjacent ports (e.g., ports 1 and 3 in Fig. 1) does not result from putting the corresponding per unit length immittance parameters equal to zero (i.e., $z_{13}=y_{13}=0$). The use of such a three-line system as a six-port reflectometer is analyzed in terms of the derived scattering parameters. The reflectometer discussed in this paper allows an unknown impedance to be measured using a standard impedance.**

I. Introduction

THE properties of coupled multiconductor systems have been extensively investigated both for homogeneous [1]–[3] and inhomogeneous [4]–[7] media. Most of the introduced analyses were based on the use of either the capacitance or the immittance matrix of the system. In many applications, e.g., analysis of couplers and reflectometers, the use of the scattering parameters of the system gives a more physical insight into the problem. The scattering parameters were used only partly for the analysis of a three-line coupler [5]. In this paper a more detailed analysis of a symmetrical three-line system (Fig. 1), which was analyzed in terms of the per unit length immittances [6], is presented in terms of the system's scattering parameters. These scattering parameters are derived in Section II. In Section III it is shown that the condition for isolation of nonadjacent ports, e.g., ports 1 and 3 in Fig. 1, is not met by putting the corresponding per unit length immittances equal to zero, i.e., $z_{13}=y_{13}=0$. In Section IV the necessary conditions allowing the use of the three-line system as one class of six-port reflectometers are derived making use of the results of the previous sections. In Section V an investigation is carried out to find the most

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R. J. Collier is with the Electronics Laboratories, University of Kent at Canterbury, The University, Canterbury, Kent CT2 7NT, England.
N. A. El-Deeb is with the Electronics Laboratories, University of Kent at Canterbury, The University, Canterbury, Kent CT2 7NT, England, on leave from The Military Technical College, Cairo, Egypt.

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