$$\vec{E}^r \propto F_r'(\omega t - \vec{k}^r \cdot \vec{r})$$

$$= F_r \left(t - \frac{\hat{k}^r \cdot \vec{r}}{c} \right) = F_r \left(t - \frac{x \sin \theta_i}{c} + \frac{z \cos \theta_i}{c} \right). \quad (\text{A4})$$

Thus, a derivative with respect to x can be replaced by a derivative with respect to t , i.e.,

$$\frac{\partial}{\partial x} \rightarrow -\frac{\sin \theta_i}{c} \frac{\partial}{\partial t}. \quad (\text{A5})$$

Substitution of this relation into (A1) yields

$$\frac{\partial E_y}{\partial z} = \mu \frac{\partial H_x}{\partial t} \quad (\text{A6a})$$

$$-\frac{\sin \theta_i}{c} \frac{\partial E_y}{\partial t} = -\mu \frac{\partial H_z}{\partial t} \quad (\text{A6b})$$

$$\frac{\partial H_x}{\partial z} + \frac{\sin \theta_i}{c} \frac{\partial H_z}{\partial t} = \varepsilon \frac{\partial E_y}{\partial t}. \quad (\text{A6c})$$

They can be further simplified as

$$\frac{\partial E_y}{\partial z} = \mu \frac{\partial H_x}{\partial t} \quad (\text{A7a})$$

$$\frac{\partial H_x}{\partial z} = \varepsilon \cos^2 \theta_i \frac{\partial E_y}{\partial t}. \quad (\text{A7b})$$

By defining

$$t' = t / \cos \theta_i \quad (\text{A8a})$$

$$H'_x = H_x / \cos \theta_i \quad (\text{A8b})$$

we finally derive the governing equations as

$$\frac{\partial E_y}{\partial z} = \mu \frac{\partial H'_x}{\partial t'} \quad (\text{A9a})$$

$$\frac{\partial H'_x}{\partial z} = \varepsilon \frac{\partial E_y}{\partial t'} \quad (\text{A9b})$$

which is a pair of one-dimensional Maxwell curl equations, and thus, the standard one-dimensional FDTD code consequently can be used for time-domain solutions. Note that H_z does not appear in the final time-domain equations; however, it may be calculated by (A6b) if necessary.

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2.4 GHz CMOS DOUBLE BALANCED GILBERT CELL MICROMIXER

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Received 28 August 2000

ABSTRACT: A double-balanced micromixer is implemented with a 0.35 μm CMOS process. The micromixer has a good frequency response, and thus only needs a small on-chip inductor for RF matching. $P_{1dB} = -10$ dBm, $IIP_3 = 0$ dBm, and a single-ended conversion loss of 4.7 dB are achieved at 2.4 GHz for a supply voltage of 3.3 V. © 2001 John Wiley & Sons, Inc. Microwave Opt Technol Lett 28: 251–253, 2001.

Key words: mixer; Gilbert cell; micromixer; RFIC; CMOS

1. INTRODUCTION

Micromixer proposed by Gilbert [1] is the ideal circuit for an RF high-frequency mixer. The advances in submicron technology have made an ultra-high-speed CMOS RFIC possible. In this paper, a CMOS version micromixer is demonstrated at 2.4 GHz for the first time. The circuit schematic is shown in Figure 1. A single-to-differential stage is constructed with $M1$, $M2$, $M3$ and two resistors. The common-gate-biased $M1$ and common-source-biased $M3$ provide equal, but out-of-phase transconductance gain when $M2$ and $M3$ are connected as a current mirror. The common-gate configuration possesses a good frequency response, while the speed of the common-source-configured $M3$ is improved drastically by adding the low-impedance diode-connected $M2$ at the input of the common-source-configured $M3$. Thus, the single-to-differential stage in Figure 1 is suitable for high-frequency operation.

2. CIRCUIT DESIGN

The single-to-differential method is not only used to turn a single signal into two balanced signals, but also to facilitate the input RF impedance matching [1, 2]. The resistance looking into point P is equal to the parallel combination of the resistance seen in the up branch and the resistance seen in the down branch. For the up branch, the resistance is equal to 33Ω , the resistance series connected with the source of $M1$, plus the inverse of $Gm1$, the transconductance of $M1$. For the down branch, the resistance is equal to 33Ω , the resistance series connected to the drain of $M2$, plus the

Contract grant sponsor: National Science Council
Contract grant number: NSC 89-2213-E-005-053
Contract grant sponsor: Ministry of Education
Contract grant number: 89-E-FA06-2-4

Figure 1 Schematic of the double-balanced CMOS micromixer

inverse of Gm_2 , the transconductance of M_2 . We can make the resistance seen at point P be $50\ \Omega$ by biasing M_1 , M_2 , and M_3 under the condition that the three transistors all have a transconductance value close to $15\ \text{mS}$. A series- L shunt- C matching circuit is used to further extend the input RF matching to higher frequencies where transistor capacitances can play an important role. Only a small inductor is needed. The RF input matching is implemented with a square spiral inductor of $1.84\ \text{nH}$ and an MIM capacitor of $0.43\ \text{pF}$. This matching policy is truly good for silicon RF design because a smaller inductor saves space in the valued chip area, and it also can be more accurately modeled. An important advantage worth mentioning is that a smaller inductor occupies less area, and also means smaller energy dissipation through substrate loss. The inductor occupies an area about of $100\ \mu\text{m} \times 100\ \mu\text{m}$, and the capacitor occupies an area of $92\ \mu\text{m} \times 92\ \mu\text{m}$.

A simple method called current-injection bias is applied to enhance the conversion gain of the mixer. A current source, which is also a diode-connected transistor M_{ci} as shown in Figure 1, is added as another branch current, which can supply part of the dc current of the drain current of the single-to-differential stage. In other words, the tail current of the differential pair in the LO stage combined with the current flowing in the added branch is the dc current of the drain current of the single-to-differential stage. By making

Figure 2 Layout with testing diagram of the double-balanced CMOS micromixer

the current of the added branch larger, we can lower the dc current of the LO differential pair and, further, make the load resistor of the LO stage larger. Because the load resistor is related to the conversion gain of the mixer, by raising the value of the load resistor, we can make the conversion gain higher [3].

3. RESULTS AND DISCUSSIONS

The mixer is implemented with a $0.35\ \mu\text{m}$ 1P4M CMOS process, and the layout with a testing diagram is shown in Figure 2. The on-wafer measurement is performed with a $2.4\ \text{GHz}$ RF signal and a $2.25\ \text{GHz}$ LO signal. A single-ended IF output is used for the measurements. The supply voltage is $3.3\ \text{V}$, and the single-to-differential stage consumes $7\ \text{mA}$. Figure 3 illustrates the conversion gain, LO-IF isolation, and LO-RF isolation as a function of LO power. The conversion loss is $4.7\ \text{dB}$ for a single-ended output when the LO power is about $-6\ \text{dBm}$. It is expected that a conversion gain of about $1.3\ \text{dB}$ can be achieved if the output is taken differentially. The LO-to-RF isolation is under $-36\ \text{dB}$. However, the isolation between the LO and IF is not good because the measurement is performed with a single-ended output. If the measurement is performed with a differential output, the LO-to-IF isolation should be better than the present measured data. The conversion gain and RF-IF isolation as a function of RF input power when the LO power is $-6\ \text{dBm}$ are illustrated in Figure 4. The $P_{1\ \text{dB}}$ is $-10\ \text{dBm}$, and the

Figure 3 Conversion loss, LO-RF isolation, and LO-IF isolation of the CMOS micromixer as a function of LO power

Figure 4 Conversion loss and RF-IF isolation of the CMOS micromixer as a function of RF power

Figure 5 Measurement of IIP3 of the CMOS micromixer

RF-IF isolation is better than 24 dB for RF power below a gain-compressed point. Finally, a two-tone intermodulation measurement is performed, and the results are shown in Figure 5. $IIP_3 = 0$ dBm is obtained from the figure.

4. CONCLUSION

Gilbert cell variant topology is demonstrated in a double-balanced mixer. With a single-ended IF output measurement, a conversion loss of 4.7 dB is achieved. It is found that the linearity of this mixer is very good: the $P_{1\text{ dB}}$ is -10 dBm and IIP_3 is 0 dBm.

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A TIME-DOMAIN BEAM-PROPAGATION METHOD FOR ANALYZING PULSED OPTICAL BEAMS IN SECOND-ORDER NONLINEAR WAVEGUIDES

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ABSTRACT: The linear time-domain beam propagation method (TD-BPM) has been extended to model the propagation of pulsed optical beams in second-order nonlinear optical material of integrated waveguides. The coupled nonlinear wave equations have been derived and discretized using the explicit finite-difference method. The nonlinear TD-BPM method developed in the present work is very efficient, and is simple to implement. © 2001 John Wiley & Sons, Inc. *Microwave Opt Technol Lett* 28: 253–257, 2001.

Key words: beam propagation; optical beams; nonlinear waveguides

1. INTRODUCTION

In the past three decades, the second-order nonlinear optical phenomenon of $\chi^{(2)}$ played a very important role in many optical applications. It was used predominantly in frequency conversion such as second-harmonic generation, parametric amplification, sum-frequency generation, and difference-frequency generation [1–4]. Most of the research focused on efficient techniques to convert to other harmonics, and recently on other effects like nonlinear phase shift to the fundamental beam [5]. At low conversion, and off phase matching between the fundamental and the second harmonic, the nonlinear phase shift behaves similar to the Kerr effect of $\chi^{(3)}$ [5]. Lately, this nonlinear phase shift process has received much attention experimentally and theoretically to achieve all-optical switching in integrated optics [6]. In order to fully understand the complicated behavior taking place in second-order nonlinear interaction, accurate and efficient techniques must be available to model such devices. On the other hand, modeling nonlinear integrated optical waveguides is difficult using analytical techniques like the coupled-mode theory, and can be more difficult if the integrated optical circuits contain multiple waveguides that have geometrical and/or material changes in all three spatial directions. Other numerical techniques like the beam-propagation method (BPM) and the finite-difference time-domain (FDTD) method are more suited for such problems. Both of these techniques were used to model, in the CW domain, optical waveguides that contain second-order nonlinear effects [7–10]. In principle, the FDTD in [9–10] should be able to model pulsed optical beams in $\chi^{(2)}$ material; however, the approximation involved showed that it is limited in its present form. The CW version reported showed limitations on the input power and the overall stability of the algorithm. In addition, the FDTD requires enormous computer resources, even for analyzing simple 2-D optical waveguides [11–12]. Recently, the beam-propagation method (BPM) was extended, using the explicit finite-difference technique, to the time domain to model pulsed optical beams [11–12]. The accuracy of the TD-BPM was tested rigorously in three different problems of homogeneous medium, metallic and dielectric waveguides. In addition to its validity in homogeneous and metallic problems, it was also concluded that the technique is well suited to study pulsed optical beams in dielectric waveguides over long distances with high efficiency.

In this work, we extend the TD-BPM [11–12] to model pulsed optical beams in the presence of second-order nonlinear effects. The parabolic coupled nonlinear wave equations for the fundamental and the second-harmonic fields were derived in the time domain. All features of time and three-dimensional spatial variations were retained, and the fundamental pulse is allowed to deplete. Previous attempts to model pulsed optical beams in $\chi^{(2)}$ material involved a 1-D (plane-wave) approximation, with many time derivative terms removed to simplify the analysis [1–4, 13–14]. In this work, the explicit finite-difference (EFD) technique was used to discretize the coupled system of nonlinear equations. The EFD method is well known for its simplicity, efficiency, and is also suited for parallel computer implementations that can model large 3-D optical devices [15]. For simplicity, we refer to the new nonlinear time-domain BPM method as TD-BPM-SHG. Finally, the method is applied to model