MILLIMETER-WAVE INTEGRATED CIRCUITS

M. V. Schneider
Bell Laboratories
Crawford Hill Laboratory
Holmdel, N. J. 07733

Abstract

Recent advances in microstrip circuit technology and solid-state device technology have resulted in hybrid integrated circuits which look attractive for use in millimeter-wave communication systems. This paper describes the design and performance of frequency converters, solid-state sources, circulators and injection-locked amplifiers at millimeter wavelengths.

Introduction

The use of integrated circuits in millimeter-wave communication systems allows a reduction in the cost and the complexity of many circuits which have been previously built with waveguide components. Integrated circuits and techniques which are suitable for systems up to 15 GHz have been described by Sobol,[1] van Heuven and van Nie,[2] and Caulton[3]. Circuits at higher frequencies can be built by modifying such techniques and by using some novel approaches which are described in this paper. We have found that millimeter-wave integrated circuits with good performance can be built by proceeding as follows.

(1) A scaled model of the circuit is built at a low frequency (1-3 GHz). The model is optimized and subsequently reduced in size to the desired millimeter-wave frequency.

(2) A substrate with a relatively low dielectric constant such as fused quartz ($\varepsilon_r = 3.8$) is used in order to obtain microstrip and stripline circuits with low loss and with dimensions which can be accurately reproduced at millimeter-wave frequencies.

(3) The circuit is mounted in a relatively narrow channel to suppress propagation of undesired higher-order modes.

These three basic design principles are discussed in more detail in the following two sections. Representative examples of specific circuits are presented in the final section of this paper.

Frequency Scaling

Millimeter-wave integration circuits can be built by optimizing the performance of a low frequency model and by scaling the circuit dimensions and device parameters by a proper factor. Although this method is well known, it has not been widely used because device parameters are difficult to scale and also because high dielectric constant materials such as alumina are expensive and difficult to obtain in large sizes. These problems can be solved using a modified device package at the low frequency and building the circuit at a frequency of the order of 1 GHz on rough and inexpensive quartz plates with an rms surface roughness of approximately 2 micrometers (80 microinches). In order to obtain circuits with similar characteristics at different frequencies, each with a characteristic length L, an RF frequency ω, a conductivity σ, a permeability μ, and a permittivity ε, it is necessary and sufficient that

$$\mu\varepsilon(\omega L)^2 = K_1 \tag{1}$$

$$\mu\varepsilon\sigma\omega L^2 = K_2 \tag{2}$$

where K_1 and K_2 are constants. Equation (1) can be readily satisfied by using a substrate which has the same properties at the low frequency and the high frequency, and by reducing all circuit dimensions at the high frequency by the same reduction factor. Equation (2) cannot be so easily satisfied. If one uses metal conductors with the same conductivity at the low frequency and the high frequency one observes an increased circuit loss at the high frequency which is given by the square root of the frequency ratio. It should also be noted that the substrate may become lossy in the millimeter-wave frequency range. The intrinsic dielectric quality factor of the line, Q_D, is given by

$$Q_D = \frac{1}{q \tan \delta} \tag{3}$$

where q is the filling factor of the strip transmission line and $\tan \delta$ the dielectric loss tangent. The loss tangent has to be determined experimentally and the filling factor can be computed from the equation[4]

$$q = \frac{\varepsilon_r}{\varepsilon_{eff}} \frac{\partial \varepsilon_{eff}}{\partial \varepsilon_r} \tag{4}$$

where ε_r is the relative dielectric constant of the substrate, and ε_{eff} the effective dielectric constant which can be derived from available computer programs such as MSTRIP.[5] One finds that the intrinsic quality factor of silica substrates in the millimeter-wave region can be as low as $3 \cdot 10^3$ due to the inclusion of OH-radicals and other contaminants. The quality factor of the microstrip line conductor is substantially lower, e.g. the Q at 30 GHz is 600 for a 50 ohm microstrip gold conductor on a 0.5 mm thick silica substrate, and the Q at 300 GHz is about 200 for the linearly scaled strip on a substrate with a thickness of 0.05 mm.

Dispersion and Higher-Order Modes

Microwave integrated circuits are usually built on substrates with a thickness of a few percent of the vacuum wavelength. Millimeter-wave integrated circuits, on the other hand, have to be fabricated on substrates which may be as thick as 10% of the vacuum wavelength in order to make the circuit losses as small as possible. The dispersive properties of line elements are therefore more pronounced in the millimeter range and must be carefully considered in the circuit design. The dispersion can be measured on a linearly scaled model at a lower frequency or it can be calculated from closed form expressions derived by Getsinger.[6] A complete mode analysis which gives the dispersion and also the cutoff frequency for higher-order modes for fully shielded lines has been made by Mittra and Itoh.[7] For most applications it is sufficient to use a simple approximation for the first higher-order mode cutoff frequency, f_c, of a microstrip, namely

$$f_c = \frac{c}{2a} \sqrt{1 - \frac{h(\varepsilon_r - 1)}{b\varepsilon_r}} \; . \tag{5}$$

Here it is assumed that the microstrip is enclosed in a rectangular shield of width a and height b. A channel width a = 4 mm and a channel height b = 2.5 mm has been used at 30 GHz for microstrip circuits on 0.34 mm thick silica substrates. The corresponding cutoff frequency is 35 GHz. A substantially wider channel could be used if a mode suppressor is built into the side wall or the top wall, or if h is substantially decreased and ε_r is increased. The first method is acceptable if the circuit operates in a frequency region which is limited by the stopband of the suppressor. The second technique is usually not desirable because the higher current density in the microstrip conductor increases the losses and decreases the circuit performance.

Performance of Specific Circuits

A downconverter which was built by following the concepts outlined above is shown in Fig. 1. The circuit which was devised by Snell and Glance[8] consists of an input RF filter, a pump filter, a low-pass IF filter and an IMPATT oscillator with its biasing circuit. A beam-leaded Schottky barrier diode is used as a mixer. The signal frequency is 30 GHz and the IF frequency is 1.6 GHz. The measured single sideband noise figure including a 0.8 dB contribution of a low-noise IF parametric amplifier is 5.5 dB.

The same downconverter was also built at 0.8 GHz, 5.3 GHz and 10.8 GHz with noise figures of 2.1 dB, 3.8 dB and 4.3 dB respectively as shown in Fig. 2. The noise figure is found to increase almost linearly with frequency by about 0.5 dB per octave. This receiver can be readily scaled to 100 GHz with a projected downconverter noise figure of about 6 dB.

A stripline injection-locked amplifier for use in a digital communications transmitter at 30 GHz is shown in Fig. 3. The circuit, designed by Trambarulo and Glance[9], consists of a stripline circulator and an IMPATT oscillator in a Y-shaped channel. The ports of the amplifier are provided with wide band transitions from stripline to waveguide for test purposes only. These transitions are not necessary upon further integration of the amplifier. The performance of a stripline circulator alone including the transitions to waveguide is shown in Fig. 4. The circulator has a 15% bandwidth for 20 dB isolation and an insertion loss of 0.25 dB. The total insertion loss including two transitions to RG-96/U waveguide is 0.55 dB.

The locking gain of the amplifier is plotted in Fig. 5 as a function of the locking bandwidth. The locking bandwidth is 380 MHz for a gain of 20 dB; the output power of the amplifier is 50 mW at 29.5 GHz. It should be possible to achieve higher output power by using an improved diode in a slightly modified circuit. In a more recent experiment an IMPATT diode mounted in the side wall of the channel with a microstrip circuit has given a CW output power of 250 mW with a corresponding RF efficiency of 5%.

Conclusions

It is shown that microstrip and stripline circuits have good performance in the millimeter range if the circuits are optimized at a lower frequency and if they are linearly scaled to the final design frequency. Frequency converters, oscillators, circulators and amplifiers have been built and tested at 30 GHz. They can be readily scaled to higher frequencies in the millimeter range.

Acknowledgments

The author thanks his colleagues B. S. Glance, W. W. Snell and R. F. Trambarulo for helpful comments and many discussions. The expert assistance of R. D'Angelo and S. R. Shah in building advanced circuits is also gratefully acknowledged.

References

1. H. Sobol, "Applications of Integrated Circuit Technology to Microwave Frequencies," Proc. IEEE, vol. 59, August 1971, pp. 1200-1211.

2. J. H. C. van Heuven and A. G. van Nie, "Microwave Integrated Circuits," Philips Technical Review, vol. 32, 1971, pp. 292-304.

3. M. Caulton, "Film Technology in Microwave Integrated Circuits," Proc. IEEE, vol. 59, October 1971, pp. 1481-1489.

4. M. V. Schneider, "Dielectric Loss in Integrated Microwave Circuits," B.S.T.J., vol. 48, September 1969, pp. 2325-2332.

5. T. G. Bryant and J. A. Weiss, "MSTRIP (Parameters of Microstrip)," IEEE Trans. Microwave Theory Tech., vol. MTT-19, April 1971, pp. 418-419.

6. W. J. Getsinger, "Microstrip Dispersion Model," IEEE Trans. Microwave Theory Tech., vol. MTT-21, January 1973, pp. 34-39.

7. R. Mittra and T. Itoh, "A New Technique for the Analysis of the Dispersion Characteristics of Microstrip Lines," IEEE Trans. Microwave Theory Tech., vol. MTT-19, January 1971, pp. 47-56.

8. W. W. Snell, Jr., "Microstrip Downconverter from 30 GHz to 1.6 GHz," W. W. Snell, Jr. and B. S. Glance, "Low-Noise Integrated Millimeter-Wave Receiver," to be published.

9. R. F. Trambarulo and B. S. Glance, "A Stripline Injection-Locked Amplifier for Use at 30 GHz," R. F. Trambarulo, "Stripline Circulators," to be published.

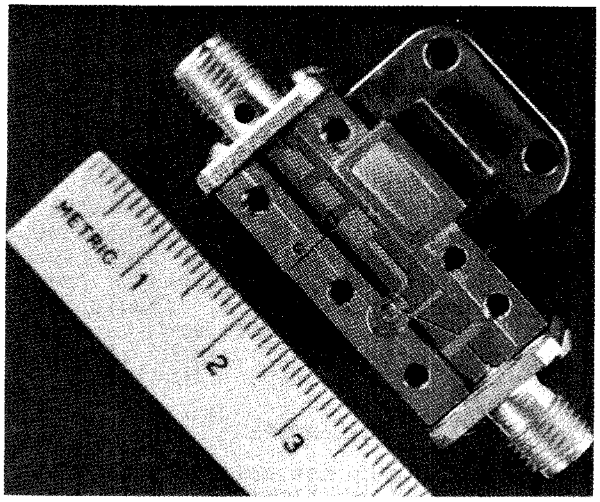

FIG. 1 TOP VIEW OF HYBRID INTEGRATED DOWN-CONVERTER FROM 30 GHz TO 1.6 GHz IN A RECTANGULAR CHANNEL WITH A WIDTH OF 4.0 mm AND A DEPTH OF 2.5 mm. THE SILICA SUBSTRATE THICKNESS IS 0.34 mm. A PROBE WHICH EXTENDS INTO AN ADJACENT RG-96/U WAVEGUIDE IS USED TO COUPLE TO THE SIGNAL.

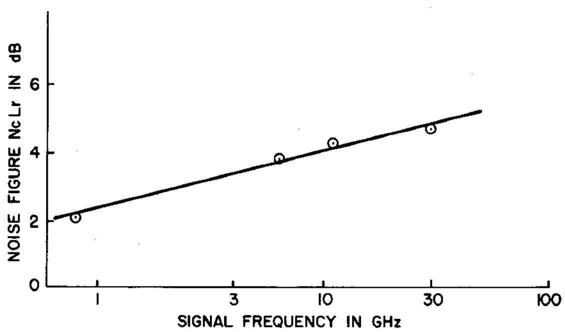

FIG. 2 NOISE FIGURE $L_C N_R$ IN DB OF LINEARLY SCALED MICROSTRIP DOWNCONVERTERS. L_C CONVERSION LOSS, N_R NOISE TEMPERATURE RATIO. THE TOTAL SINGLE SIDEBAND NOISE FIGURE AT 30 GHz INCLUDING A 0.8 DB CONTRIBUTION OF A LOW-NOISE IF AMPLIFIER IS 5.5 DB.

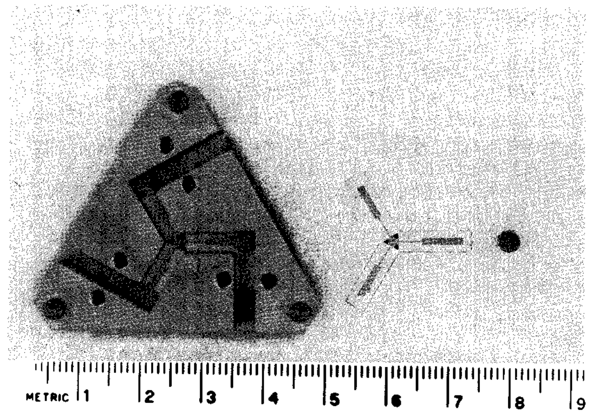

FIG. 3 TOP VIEW OF INJECTION-LOCKED AMPLIFIER BUILT IN STRIPLINE. A Y-SHAPED CHANNEL IS USED WITH TRANSITIONS TO RG-96/U WAVEGUIDE. THE IMPATT DIODE IS SHOWN ON THE FAR RIGHT.

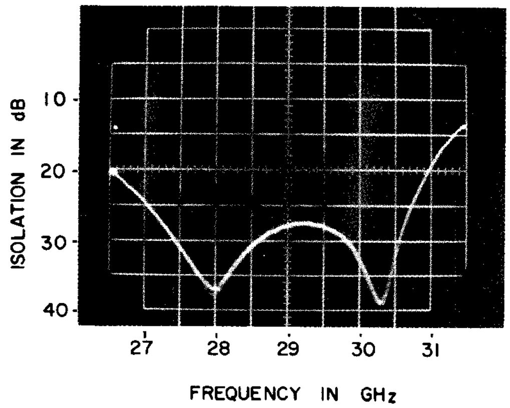

FIG. 4 ISOLATION OF 30 GHz STRIPLINE CIRCULATOR. THE BIASING FIELD FOR THE MAGNESIUM FERRITE WITH A SATURATION MAGNETIZATION OF 2,800 GAUSS IS 1,600 OERSTEDS.

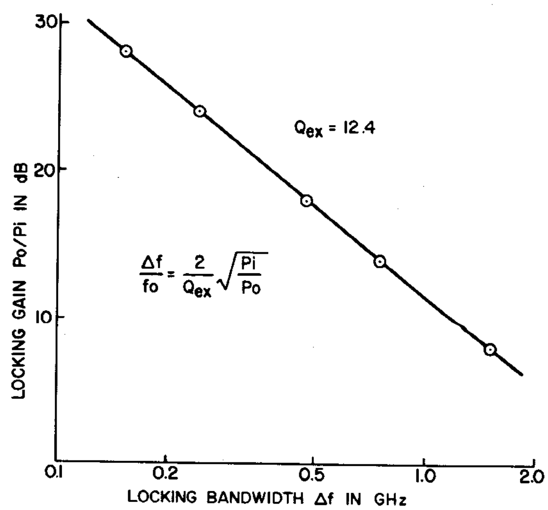

FIG. 5 LOCKING CHARACTERISTICS OF INJECTION-LOCKED AMPLIFIER AT 30 GHz.