packaged such a device on a TO can and used a constant current of 6mA (100 μ W output power), corresponding to a current density of 10kA/cm² (9 μ m diameter circle). The laser was randomly selected from the low resistance lasers, with a threshold current of 2mA at 15°C. After 4000h, there are no signs of degradation in the L-I characteristics of this laser. Since this experiment was carried out on a single device, this cannot be interpreted as proof of reliability. Accelerated aging with many devices must be carried out before any definitive statement about long term reliability can be made.

Conclusion: Long wavelength vertical cavity lasers have developed significantly in recent years. With pulsed operation temperatures as high as 124°C achieved it seems very likely that 85°C CW operation will soon be attained. At this point practical considerations for commercial devices will become important. These include simple packaging as well as higher output powers. In addition, device reliability will be one of the critical issues to address. 4000h operation is the first reported aging time of any long wavelength vertical cavity laser.

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Using a forced nonlinear oscillator to implement a QPSK modulator without a 90 degree phase shifter

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A new quadriphase-shift keying (QPSK) modulator is presented. It comprises two nonlinear oscillators, one RC differentiator and two voltage control current sources (VCCSs); it is named the 'direct QPSK modulator'. Experimental results are also reported. Such a modulator is suitable for generating direct sequence spread spectrum signal in a wireless communication system, and can be packaged into an IC.

Introduction: Owing to the spectrum efficiency and the constant envelope, QPSK modulation is widely used in various kinds of communication systems. Traditionally, a double balance modulator and a 90° phase shifter are usually employed to implement a QPSK modulator. In this Letter, we propose a new architecture for realising such a modulator. The kernel of this circuit comprises

two nonlinear oscillators, one RC differentiator, and two voltage control current sources (VCCSs); it does not need a phase shifter or a balance modulator. In the following, we will show that the implementation of this circuit is very easy, and that the performance is acceptable.

Fig. 1 Structure and principle of proposed QPSK modulator

- a Forced response of ideal LC resonator
b Nonlinear resistor realisation
c Block diagram of QPSK modulator
d Phase-plane trajectories of forced oscillator

Principle: When a series of current pulses are used to excite an ideal LC resonator, the output voltage will be a BPSK signal [1]. Similarly, if the exciting signal is a series of current impulses then the response voltage will also be a BPSK signal. Fig. 1a shows that the carrier of the former BPSK signal is orthogonal to the carrier of the latter BPSK signal. Owing to the inherent loss of an LC resonator, the output voltage will gradually decay to zero. Therefore, if the ratio of bit rate to the carrier frequency (R_b/f_c) is too low, the output voltage will no longer be a proper BPSK signal. In [1], a comparator is used to compensate this effect. In this Letter, we will use a nonlinear resistor, as shown in Fig. 1b, to eliminate this effect, this circuit then becomes a nonlinear oscillator. The characteristics of our nonlinear resistor are shown as the solid line in Fig. 1b. If $|V| \leq A$, the resistor is an active device, so the oscillation amplitude will grow; otherwise, the amplitude will decay. Thus the oscillation amplitude will be limited to a constant value. If the driving-point characteristics is a 'cubic' function (dotted line), the circuit will be the well-known Van der Pol oscillator [2, 3].

Both [2, 3] point out that a limiting cycle is a unique feature of a nonlinear oscillator, and becomes an ellipse if the nonlinearity is small. We know that the projection of an elliptic trajectory on the V -axis is a sinusoidal wave. If we use sufficient external force to change the state of this oscillator at the right time and right place, the projection will be a BPSK signal. This idea is shown in Fig. 1d, and it depicts two kinds of forced trajectory of different BPSK signals.

Employing the concept discussed above, we can design a novel QPSK modulator as shown in Fig. 1c. The series-to-parallel multiplexer transfers the input bit stream to the I and Q channels alternately. The Q channel information is injected to the VCCS directly; the I channel information is passed through a differentiator to generate the impulse signal, and this signal is then injected to the VCCS. The output current of the VCCS is used to excite the nonlinear oscillator following it to generate the desired BPSK signal. Finally the QPSK signal is obtained by summing these two BPSK signals.

Experimental results: Fig. 2a shows the Q channel modulating-bit-stream and its relative BPSK signal. Similarly, Fig. 2b depicts the I channel modulating-bit-stream and its corresponding BPSK signal. In this experiment, carrier frequency and bit rate are 5MHz and 100kbit/s, respectively. It is obvious from the Figure that the transition of the bit-stream will reverse the carrier phase. This phenomenon corresponds to the state change shown in Fig. 1d. From this figure we can also see that the oscillation voltage is a constant value and the $R_b/f_c = 0.02$. We use a digital transmission analyser (Anritsu ME520A) to generate a PN sequence $2^{10}-1$ long as the

modulating-bit-stream, and a vector signal analyser (HP89441) to appraise the performance of this QPSK modulator. The bit rate and carrier frequency are now 2.5Mbit/s and 5MHz, respectively.

Fig. 2 Resultant output waveforms

a I channel modulating bit stream and corresponding BPSK signal
b Q channel modulating bit stream and corresponding BPSK signal

Fig. 3 Spectrum of QPSK signal and demodulation output

a Signal constellation
b Modulator output spectrum
c Eye diagram of in-phase channel
d Eye diagram of quadrature-phase channel

The signal constellation and spectrum of this QPSK modulator are shown in Fig. 3a and b. Fig. 3c and d shows the eye-diagrams of demodulated bit streams corresponding to the I and Q channel, respectively. It is obvious that such a modulator can indeed generate a QPSK signal, and that the performance is acceptable.

Conclusion: A new type of QPSK modulator comprising a nonlinear oscillator, differentiator, and a VCCS has been implemented. It can generate the correct QPSK signal, and the ratio of bit rate to carrier frequency R_b/f_c may be very small since the output signal level decay can be compensated using the proper nonlinear resistance. The realisation of this modulator is fairly simple and its performance is acceptable. Such a modulator is suitable in practical communication systems provided we can lock its carrier frequency to a reference signal.

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Determination of W-band noise parameters

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A computer controlled fully automated noise measurement system and method where correlation matrices are used to de-embed the noise parameters of the test device is described. The first reported measured noise parameters and the respective noise parameter uncertainties of a passive ambient device operating at 80GHz are presented. These results show very good agreement with the noise parameters derived from the scattering parameters of the passive ambient device.

Introduction: Accurate noise parameter measurements of active devices are required due to the continuing technological advances in monolithic millimetre-wave integrated circuits. A method for the determination of noise parameters at W-band frequencies is presented. To verify the noise measurement system, a passive ambient device in the form of a waveguide filter was measured. Although these measurements were performed in a waveguide, this method could equally be applied to on-wafer measurements. The measured and calculated noise parameters [1] and the uncertainties at 80GHz for the waveguide filter are presented.

Fig. 1 W-band noise measurement system

Noise measurement method: A diagram of the automated measurement system is shown schematically in Fig. 1. It allows measurement of noise and scattering parameters of a device under test (DUT) without any reconnection. A correlation matrix technique [2, 3] is used which provides a convenient means of de-embedding the noise parameters of the DUT from those of the measurement system. Using the notation in Fig. 1, the measurement system can be treated as a cascade of stages where the correlation matrices $[C_{DUT}]$ and $[C_{REC}]$ describe the noise contribution of the DUT and the receiver, respectively. The scattering parameters of the DUT are represented by the transmission matrix $[A_{DUT}]$. Consequently, the total noise in the system can be described by the correlation matrix $[C_T]$:

$$[C_T] = [C_{DUT}] + [A_{DUT}][C_{REC}][A_{DUT}]^+ \quad (1)$$

where the '+' is used to denote Hermitian conjugation.

The method used to determine the noise parameters of the receiver is based on a technique described by Adamian and Uhler [4]. In this procedure, the noise source is only used in the 'on' and 'off' states (corresponding to the noise source temperatures T_n and T_a , where T_a is ambient temperature), solely to determine the gain-bandwidth of the receiver, kGB , where k is the Boltzman constant. For the present measurements it is assumed that the ambient temperature T_a is equal to the standard noise temperature $T_o = 290K$, although another value could be used if appropriate.

With the noise source set to its 'off' state, the noise power is measured for each of a randomly chosen set of impedance points synthesised by the automated tuner [5]. The noise factor of the receiver for each source impedance point is then determined from

$$F_{REC}(\Gamma_s) = \frac{P_{meas}}{T_o M_{meas}} \left[\frac{1}{kGB} \right] - \frac{T_{s1}}{T_o} + 1 \quad (2)$$