GaAs monolithic 1.5 to 2.8 GHz tunable ring oscillator with accurate quadrature outputs

Wen-Chieh Wu and Hao-Hsiung Lin

The design, fabrication and characteristics of a 1.5 to 2.8 GHz tunable ring oscillator with two quadrature outputs are described. Tuning range is 1.3 GHz or 60.5% bandwidth. The phase difference between two outputs is proved to be 90°. The oscillator phase noise is −81.27 dBc/Hz at 100 kHz offset from the carrier at 2.736 GHz.

Introduction: Circuits such as binary-phase-shift-keyed (BPSK) demodulators or quadricorrelator frequency discriminators require two quadrature inputs with accurate 90° phase difference [1]. There are limitations for the circuit, which is composed of a voltage-controlled oscillator (VCO) and a 90° phase-splitter. The frequency tuning range of the VCO is usually quite small. The 90° phase-splitter constructed by passive elements is designed only for a narrow frequency band. This problem is severe, especially when a wide frequency tuning range is required. In addition, the passive LC elements in phase-splitters are very cumbersome for modern integrated processes. To provide a wide tunable oscillator with accurate 90° quadrature outputs, a ring oscillator was selected. Since the ring oscillator is constructed by chaining inverter stages, the phase-difference between any two stages is fixed. This monolithic oscillator can provide 90° phase-difference between two outputs over a wide frequency range and no inductor is needed.

Circuit: The block diagram of the ring oscillator is shown in Fig. 1. In this design, only two GaAs differential-pair inverters are adopted. Each stage provides 90° phase-delay in one oscillating cycle. The inverter includes an active-loaded differential pair and two common drain buffer stages for two outputs. Symmetrical layout is performed and the integrated circuit was fabricated by Hexawave 1 μm GaAs MESFET MMIC low-cost process.

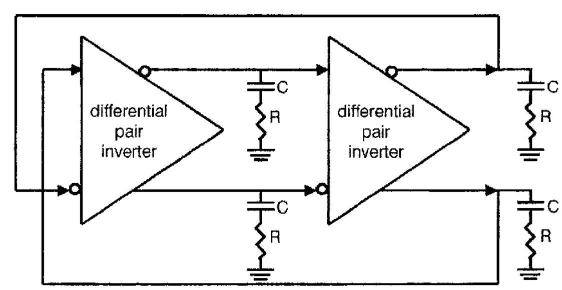

Fig. 1 *Block diagram of tunable ring oscillator*

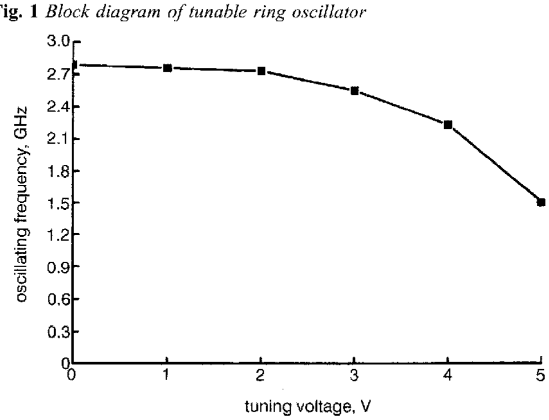

Fig. 2 *Oscillating frequency against tuning voltage*

Measurements and analysis: With tuning voltage varying from 0 to 5 V, the oscillating frequency varied from 2.8 to 1.5 GHz as shown in Fig. 2. Tuning range is 1.3 GHz or 60.5% bandwidth. By controlling the varactor diode voltage, we can easily modulate the delay time of each stage. At 2 V tuning voltage, measured phase noise of the oscillator is −81.27 dBc/Hz at 100 kHz offset from the carrier at 2.736 GHz as can be seen in Fig. 3. For this type of ring oscillator, the time jitters are approximately the same as the original differential inverters. The fully differential design helps to reduce initial jitter due to common-mode noise and especially due to power supply coupling, which is a major source of jitter in high-frequency oscillators [2, 3]. RF oscillator concepts with impedance matching design and a symmetrical layout are also helpful to minimise jitters in this digital oscillator. Parallel-connected on-chip capacitors with different values are used for power source stabilisation in the oscillating frequency. To measure the phase difference between two outputs, a two-channel high-speed oscilloscope was used. At 3 V tuning voltage, oscillating frequency is 2.55 GHz and the measured *X–Y* mode result is shown in Fig. 4. Due to amplitude imbalance, it is easy to distinguish between the horizontal long axis and the vertical short axis in the ellipse. It is obvious that the maximum absolute value of output 1 is the zero point of output 2, and vice versa. 90° phase difference between outputs is observed

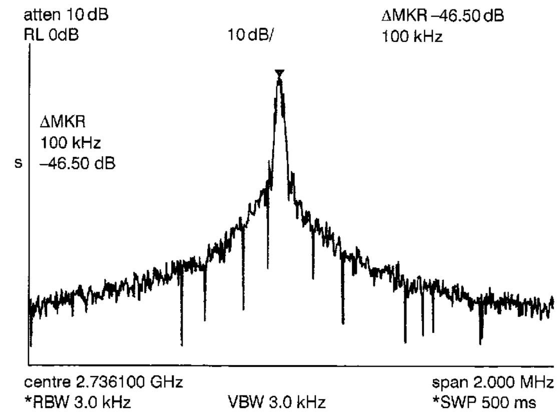

Fig. 3 *Measured phase noise*

Tuning voltage, 2 V; measured frequency, 2.736 GHz; phase noise, −81.27 dBc/Hz at 100 kHz offset

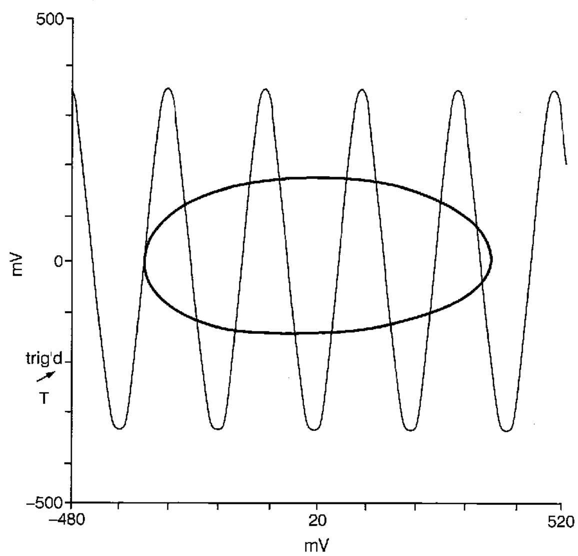

Fig. 4 *Oscillating frequency (2.55 GHz) and measured X–Y mode result (flat ellipse)*

Tuning voltage, 3 V

Conclusion: We have reported the design, fabrication, and measurement of a novel 1.5 to 2.8 GHz-wide bandwidth monolithic tunable ring oscillator with two quadrature outputs based on 1 μm GaAs MESFET MMIC low-cost process. The output phase noise is −81.27 dBc/Hz at 100 kHz offset from the carrier at 2.736 GHz. The combination of circuit simplicity, no inductor architecture, accurate 90° quadrature outputs and 1.3 GHz or 60.5% bandwidth tuning range makes this ring oscillator attractive for many applications such as BPSK demodulators or high-speed clock-recovery circuits.

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Modelock-avoiding synchronisation method

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A simple synchronisation method avoiding the modelock phenomenon has been realised using dynamic network coupling. Two-dimensional arrays of millimeter-wave power-combining and beam-scanning control systems are considered as an application of the method. The effect, limitations and robustness of the proposed method are investigated numerically.

Introduction: Networks of interconnected oscillators emerge in a wide range of engineering issues. Examples are known in millimetre-wave power-combining and beam-scanning control systems [1], a novel VLSI clocking [2], and Josephson junction arrays, where implementation of two-dimensional (2D) synchronous arrays has been a technical challenge since a number of oscillators are packed in a limited space and are required to oscillate in unison. As opposed to one-dimensional (1D) linear arrays of oscillators, planar arrays of oscillators can sometimes exhibit 2D phase-lagged stable synchronous patterns called modelock or travelling (spiral) waves. This modelock has been a notorious hazard since it hampers the desired in-phase synchronisation of the oscillator arrays.

In a 2D array of voltage-controlled oscillators (VCOs) for VLSI clocking [2], modelock is avoided by adding a special phase detector (PD) the response of which decreases monotonically beyond a phase difference of $\pi/2$ to each VCO. However, in solid-state circuits such as MESFET oscillators for millimetre-wave generation, the interaction between oscillators is due to the 'injection-locking' mechanism and the resulting synchronisation characteristics (corresponding to the PD response) come from the intrinsic nonlinearity of the oscillator. Thus, in such cases avoiding modelock by tuning the synchronising characteristics may not be straightforward (if not impossible), as opposed to the case of the VLSI clocking circuit.

In this Letter we propose an alternative synchronising method that avoids modelock by introducing dynamic coupling with only on–off switches to the array (see Fig. 1). The basic idea comes from the observation that nonregular (percolation-like) 2D networks of oscillators attain the in-phase synchronisation state by destruction of the 'core' of the spiral wave pattern (centre of the modelock). Systematic numerical simulations are carried out to consider the effectiveness of the method for possible applications to millimetre-wave power-combining and beam-scanning control systems. The limitations and robustness of the method are also investigated.

Phase dynamics in 2D oscillator arrays: We assume here a weak coupling between adjacent oscillators, and sufficiently uniform oscillator characteristics. Under such conditions, a systematic derivation of the phase equation for oscillators can be made (e.g. see [1]), which eventually takes the following form:

$$\dot{\theta}_i = \omega_i + \Delta\omega_m \sum_j \sin(\Phi + \theta_j - \theta_i) \tag{1}$$

where θ_i and ω_i represent the oscillation phase and free-running frequency of the ith oscillator, respectively. $\Delta\omega_m$ is interpreted as the locking range of each oscillator which is assumed to be small and the same for all oscillators. The phase lag Φ reflects signal delay, which cannot be neglected for the case of radiative coupling. However, if the coupling is made by one-wavelength waveguides, Φ is assumed to be 0 and we focus on this case. The above discussion is valid for any network topology. For the 2D oscillator array with switched coupling as shown in Fig. 1, (1) can be modified to the following phase equation

$$\dot{\theta}_i = \omega_i + \Delta\omega_m \sum_j S_i S_j \sin(\Phi + \theta_j - \theta_i) \tag{2}$$

where the additional parameter S_i takes $1/0$ depending on the on/off states of the switch between the ith oscillator and the coupling network. The summation $\sum_j$ is taken for adjacent (jth) oscillators to the ith oscillator. Here, we assume each switch independently takes the on/off states alternatively for a time span T_{on}/T_{off}, respectively. It is clear that if all switches are on, (2) reduced to (1). Conversely, if all switches are off, (2) becomes $\dot{\theta}_i = \omega_i$, which implies that all oscillators are free-running. Between these two extremes, we have intermediate states where any two oscillators with $S_i = 1$ are connected by the network and a certain amount of oscillators with $S_i = 0$ are disconnected from the network, forming a percolation-like, global network of interacting oscillators.

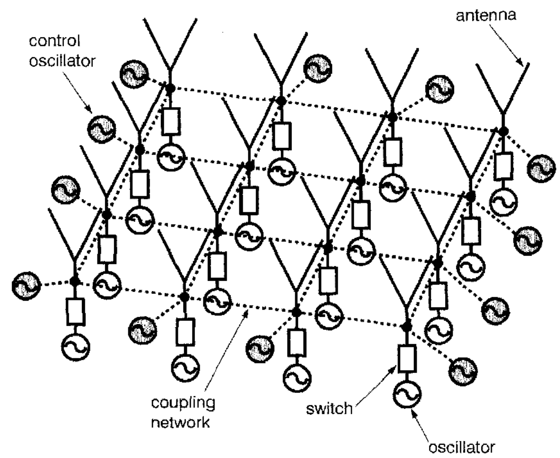

Fig. 1 *2D square array oscillators with switched couplings*

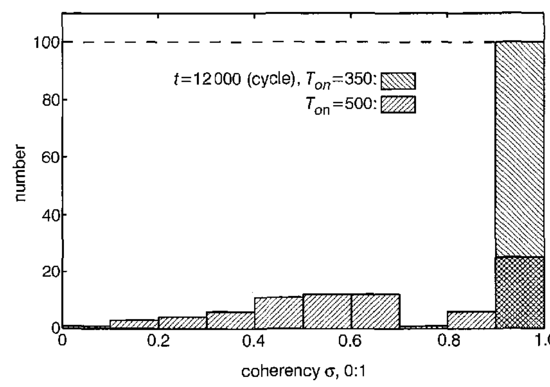

Fig. 2 *Coherency distribution of σ after 12000-cycle oscillations*

We performed numerical simulations of (2) for two cases of $(T_{on}, T_{off}) = (350, 150)$ and $(T_{on}, T_{off}) = (500, 0)$, respectively, where 100 trials are made for 30×30 oscillator array networks with random initial oscillation phasers and switching states. The natural frequencies ω_i are chosen uniformly from [0.95, 1.05] at random. To measure the degree of phase synchronisation, the phase coherency σ is introduced as follows:

$$\sigma = \sum_{j=1}^{N} S_j \exp(i\theta_j) \bigg| \bigg/ \sum_{j=1}^{N} S_j \tag{3}$$