without guiding frequency filter in loop

Fig. 2 Measured bit-error rate as function of total path, with and without guiding frequency filter in loop

- 2.5 Gbit/s
- with guiding frequency filter
- without guiding frequency filter

with a 10 GHz FWHM optical filter (to reduce spontaneous emission noise) inserted just before the detectors, yielded an error free distance (EFD) of slightly greater than 11000 km. That EFD is in accord with theoretical predictions, on the assumption that the effective phase margin of the decision circuits in the presence of significant amplitude noise is about 210 ps. (For an error rate \( < 10^{-10} \), a window at least \( \pm 6\sigma \) wide is required, where \( \sigma \) is the standard deviation of the Gordon-Haus jitter; here \( \sigma = 16 ps \) at 11000 km.)

The second set of measurements were carried out with a special, narrow-band (curvature at peak, \( 1.37 \times 10^{-10} \) GHz \(^{-2} \)) frequency filter inserted into the loop itself. The effect of that filter, as recently described elsewhere, is to guide the pulses in the frequency domain, and thus to dampen jitter in pulse arrival times from the Gordon-Haus effect. The filter also incidentally reduces the spontaneous emission spectral width, after many round trips, to little more than that of the soliton, so no additional filter is required in front of the detectors. Analysis is further complicated by the mixed effects of the filter on amplitude noise, i.e. its tendency to decrease amplitude fluctuations of the solitons, while increasing noise in the zeros. Nevertheless, the resultant increase in EFD (to slightly beyond 14000 km) corresponds well to the expected extension in distance to recover the same Gordon-Haus jitter (\( \sigma = 16 ps \)) and to the 210 ps effective phase margin. Finally, we note that both sets of measurements were almost certainly affected by excess noise at or near the signal frequency, corresponding to dispersive wave radiation shed by the initially chirped mode-locked diode laser pulses as they evolve into exact solitons. Thus, there should be room for still further increase in the EFD with an improved pulse source.

Elsewhere we have shown both theoretically and experimentally that orthogonally polarised solitons remain so over transoceanic distances, and that the interaction between them is greatly diminished. We have further shown how polarisation division multiplexing (PDM) based on those facts should allow a doubling of the bit rate with negligible increase in error rate. The results of this work thus imply that an error-free, single channel bit rate of at least 5 Gbit/s over transoceanic distances should be easily achieved. We also note that a further doubling of the rate (to \( 10 \) Gbit/s) through wavelength division multiplexing with two closely spaced (\( \Delta \lambda \leq 0.5 \) nm) channels, should be immediately practical.

Introduction: Recently, current mirrors with or without a conventional operational amplifier are used to construct many high performance current-mode circuits, such as current followers (CF), second generation current conveyors (CCII), etc. Among them, application of CCII\(^d\) and CFs\(^e\) in the design of voltage amplifiers has the merit of wide bandwidth and accurate amplifier performance.

CCII\(^d\) have also been widely employed to synthesise many filtering, immittance simulating, and oscillator circuits. However, considering the active sensitivity, the voltage tracking error of a CCH will influence many important circuit parameters (e.g. the natural frequency and the quality factor in filtering applications, the oscillation frequency in oscillator applications). Fortunately, this problem can be avoided in the CFs because of the property of virtual ground at the input. Therefore, CFs to develop various analogue signal processing circuits, such as analogue filters, amplifiers, and oscillators, seems to be remarkably promising.

We propose a new configuration for realising active RC oscillators using a CF. Using additional Cf's, a sinusoidal quadrature oscillator can be realised. The nonideal effects of a CF including nonzero-offset voltage and nonzero-offset current on the proposed oscillators are analysed. Simulation results which confirm the theoretical analysis are obtained.

References

CURRENT-MODE OSCILLATORS USING SINGLE CURRENT FOLLOWER

Indexing terms: Oscillators, Circuit theory and design, Circuit design

A new configuration for realising sinusoidal quadrature oscillators using current follower (CF) is presented. The effects of nonzero-offset voltage and current (at the input and output terminals of a current follower, respectively) on the oscillator are discussed. The resulting oscillating frequency is insensitive to the current tracking error of a CF. Simulation results which agree with theoretical analysis are obtained.

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istics of the CF, shown in Fig. 1, can be described by the following matrix equation:

\[
\begin{bmatrix}
\hat{y}_2 \\
\hat{y}_3
\end{bmatrix} = \begin{bmatrix}
\pm K & 0 \\
0 & 0
\end{bmatrix}
\begin{bmatrix}
\hat{y}_x \\
\hat{y}_y
\end{bmatrix}
\]

(1)

where the plus and minus signs of the current transfer ratio denote CF+ and CF-, respectively. Consider the circuit shown in Fig. 2. Assuming the CF- to be ideal, routine circuit analysis for unity loop-gain of current yields

\[y_f(y_3 + y_4 + y_2 + y_f) + y_2 y_3 - (K - 1)y_f y_4 = 0\]

(2)

As an example, if \(y_1 = sC_1 + 1/R_1 = sC_1 + G_1\), \(y_2 = sC_2\), \(y_3 = 0\), and \(y_4 = 1/R_4 = G_4\), as shown in Fig. 3, the following equation can be obtained:

\[s^2C_1C_2R_1R_4 + sC_2R_4 + C_1R_4 - (K - 1)C_2R_4 + 1 = 0\]

(3)

Therefore the frequency and the condition of oscillation can be described as

\[\omega_0 = \left(\frac{1}{C_1C_2R_1R_4}\right)^{1/2}\]

(4)

and

\[C_1 = C_2\left(K - 1 - \frac{R_4}{R_1}\right)\]

(5)

To obtain the quadrature sinusoidal output signals, the circuit of Fig. 4 can be constructed. By introducing two additional CFs, we can sense out the currents in resistor \(R_1\) and capacitor \(C_1\), which have 90° phase difference. Based on eqn. 2, many appropriate RC element combinations can yield useful oscillator circuits which are summarised in Table 1. The quadrature sinusoidal output signals can also be obtained in most of the configurations in Table 1.

**Performance analysis:** Consider a nonideal CF having a nonzero offset voltage \(V_{\text{off}}\) at terminal X and a nonzero offset current \(I_{\text{off}}\) at terminal Y. Let us consider the effects of such nonzero offset voltage and offset current separately.

First, let us consider the effects of the nonzero-offset current \(I_{\text{off}}\) on the circuit shown in Fig. 2. Because \(I_y = -KI_x\),

\[I_y = I_{\text{off}}\left(y_f(y_3 + y_4 + y_2)(y_3 + y_4)\right)_{s=0}\]

(6)

For the circuit of Fig. 3, the DC magnitude of the current \(I_y\) can be derived as the following:

\[I_y = I_{\text{off}}\left(KsC_1R_1\left(sC_1R_4 + 1\right) + sC_2R_4\right)_{s=0} = 0\]

(7)

**Table 1 OSCILLATORS FROM EQN. 2**

<table>
<thead>
<tr>
<th>Network</th>
<th>(y_1)</th>
<th>(y_2)</th>
<th>(y_3)</th>
<th>(y_4)</th>
<th>Condition for oscillator</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>(sC_1 + 1/R_1)</td>
<td>(sC_2)</td>
<td>0</td>
<td>(1/R_4)</td>
<td>(C_1 = C_2\left(K - 1 - \frac{R_4}{R_1}\right))</td>
</tr>
<tr>
<td></td>
<td>(sC_1 + 1/R_1)</td>
<td>(sC_2)</td>
<td>(1/R_3)</td>
<td>(1/R_4)</td>
<td>(C_2\left(\frac{1}{R_3} + \frac{1}{R_4}\right) = C_2\left(K - 1 + \frac{1}{R_1} - \frac{1}{R_3}\right))</td>
</tr>
<tr>
<td>3</td>
<td>(1/R_1)</td>
<td>(sC_2)</td>
<td>(sC_3 + 1/R_3)</td>
<td>(1/R_4)</td>
<td>(C_2\left(\frac{1}{R_2} + \frac{1}{R_4}\right) = C_2\left(K - 1 - \frac{1}{R_2} - \frac{1}{R_3}\right))</td>
</tr>
<tr>
<td>4</td>
<td>(sC_1 + 1/R_1)</td>
<td>(1/R_2)</td>
<td>(1/R_3)</td>
<td>(sC_4)</td>
<td>(C_2\left(\frac{1}{R_2} + \frac{1}{R_4}\right) = C_2\left(K - 1 - \frac{1}{R_2} - \frac{1}{R_3}\right))</td>
</tr>
<tr>
<td>5</td>
<td>(sC_1 + 1/R_1)</td>
<td>(sC_2)</td>
<td>(sC_3)</td>
<td>(1/R_4)</td>
<td>(C_2\left(\frac{1}{R_2} - \frac{1}{R_3}\right) = C_2\left(K - 1 - \frac{1}{R_2} - \frac{1}{R_3}\right))</td>
</tr>
<tr>
<td>6</td>
<td>(sC_1)</td>
<td>(1/R_2)</td>
<td>(sC_3 + 1/R_3)</td>
<td>(sC_4)</td>
<td>(C_2\left(\frac{1}{R_2} - \frac{1}{R_3}\right) = C_2\left(K - 1 - \frac{1}{R_2} - \frac{1}{R_3}\right))</td>
</tr>
<tr>
<td>7</td>
<td>(sC_1 + 1/R_1)</td>
<td>(1/R_2)</td>
<td>(sC_3)</td>
<td>(sC_4)</td>
<td>(C_2\left(\frac{1}{R_2} + \frac{1}{R_4}\right) = C_2\left(K - 1 - \frac{1}{R_2} - \frac{1}{R_3}\right))</td>
</tr>
<tr>
<td>8</td>
<td>(1/R_1)</td>
<td>(sC_2)</td>
<td>(sC_3)</td>
<td>(1/R_4)</td>
<td>(C_2\left(\frac{1}{R_2} + \frac{1}{R_4}\right) = C_2\left(K - 1 - \frac{1}{R_2} - \frac{1}{R_3}\right))</td>
</tr>
<tr>
<td>9</td>
<td>(sC_1)</td>
<td>(1/R_2)</td>
<td>(1/R_3)</td>
<td>(sC_4)</td>
<td>(C_2\left(\frac{1}{R_2} + \frac{1}{R_4}\right) = C_2\left(K - 1 - \frac{1}{R_2} - \frac{1}{R_3}\right))</td>
</tr>
</tbody>
</table>
Let us consider the effects of the nonzero-offset voltage \( V_{\text{off}} \). Eqn. 8 can be obtained as

\[
I_s = V_{\text{off}} \frac{K_jy_j(y_j + y_3) + y_2y_3}{y_j(y_j + y_3 + y_4) + y_2y_3 - (K - 1)y_4} \tag{8}
\]

The DC magnitude of the current \( I_s \) of the circuit shown in Fig. 3 can be calculated as

\[
I_s = V_{\text{off}} \frac{K_jC_j(C_j + 1)}{(C_j + 1)(C_j + 1) - (K - 1)C_j} \tag{9}
\]

From eqns. 7 and 9, the DC magnitude of the current \( I_s \) will be zero. Hence, latchup will not occur in this oscillator. The other conditions in Table 1 can be calculated by eqns. 6 and 8.

Fig. 4 Proposed quadrature sinusoidal oscillator

Sensitivity analysis and simulation results: Let the characteristic of a nonideal CF be given by

\[
i_s = n_i \bar{i}_s \tag{10}
\]

where \( n_i = K(1 - e_i) \) and \( e_i \) denotes the current tracking error of a CF. A detailed analysis for the transfer functions gives the following expressions (for Fig. 3):

\[
s^2 C_1 C_2 R_1 R_2 + s(C_1 R_1 + C_2 R_2) - K R_j (1 - e_i) + 1 = 0 \tag{11}
\]

Hence, the oscillation frequency \( \omega_0 \) will be insensitive to the current tracking error of a CF.

A CF—has been proposed with the circuit, shown in Fig. 5, and transistor sizes described in Table 2. The simulation results of the oscillator in Fig. 3 using this CF—with \( C_1 = C_2 = C, \quad R_1 = R_a \), and \( K = 3 \) are shown in Figs. 6–8.

Fig. 5 Simulation circuit for CF—

Conclusions: We have proposed new oscillators using a single CF. The nonideal effects of a CF on the proposed oscillator.
are discussed. The active sensitivities have also been analysed. The resulting oscillating frequency will be insensitive to the table current tracking error of a CF. Simulation results are given to demonstrate the feasibility.

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References


MILLIMETRE WAVE PHASE REVERSAL INTEGRATED OPTICAL MODULATOR IN FIN LINE

Indexing terms: Optical modulation, Integrated optics, Optics

A new millimetre wave Ti:LiNbO$_3$ optical modulator is demonstrated from 25-40GHz. Phase modulation of 60°/W$^{-1}$ is observed at 30GHz. The modulator is constructed in a fin line geometry. Detailed characterisation of the microwave transmission line is accomplished by application of the method of lines.

Introduction: Considerable effort has been devoted to the development of optical modulators operating well into the millimetre region. Both resonant$^1$ and broadband$^2$ Ti:LiNbO$_3$ devices have been reported with modulation frequencies in excess of 30GHz. Indeed, recently Bridges et al.$^3$ reported modulation at 60GHz, though with relatively low electro-optic efficiency. Many of the difficulties associated with high frequency design are related to the microwave properties of the modulator. In addition to the problem of electrode loss, coupling of the electrical drive signal onto the modulator electrodes becomes problematic at high frequencies (>40GHz); the excitation of undesirable modes in the microwave circuitry must be avoided. Here we demonstrate a new Ti:LiNbO$_3$ millimetre wave optical modulator which is excited directly from rectangular waveguide. With relatively minor modifications we believe a similar device could be constructed for operation at 60GHz.

Design: Fig. 1 shows a cutaway diagram of the modulator. The device consists of a slot line on a 120μm Z-cut lithium niobate substrate orientated to lie in the E plane of a WR28 waveguide. This configuration is known as a unilateral fin line. In the region of the discontinuity corresponding to the abrupt introduction of the dielectric, the structure is double moded, as can be seen from Fig. 2. Because of the symmetry of the discontinuity and the profile of the excitation field we assume that most energy is coupled into the dominant mode. Any energy that is coupled into the higher order mode through interaction of the evanescent field with the tapered region will be absorbed as the higher order mode is cut off by the tapering fins.

With the microwave energy propagating in the fundamental mode an electrode taper is used to effect a transition to the desired mode. The tapering fins.

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Any energy that is coupled into the higher order mode through interaction of the evanescent field with the tapered region will be absorbed as the higher order mode is cut off by the tapering fins.

With the microwave energy propagating in the fundamental mode an electrode taper is used to effect a transition to the interaction region. Careful design is essential to prevent excessive reflection and reduce loss (by minimising the length). The tapered transmission line can be analysed approximately providing that the propagation constant and impedance are known within the tapered region.$^2$ This design data was generated numerically using a hybrid mode analysis based on the method of lines.$^3$ Fig. 3 shows the effective index and power-