where the state vector $x(k)$ represents the output signals at the delays. It is proved that the constant-input limit cycles in this filter can be eliminated by employing the technique of controlled-rounding if

$$P = [p_1 \quad p_2 \cdots \quad p_m] = (I - A)^{-1}B$$

is machine representable. The structures given in References 1-3 are free of zero-input limit cycles, and hence for eliminating constant-input limit cycles they can be modified so that they satisfy eqn. 2. During this process, the recursive part of the filter is not altered so that the passband sensitivity property of the modified filter remains the same as that of the original filter. Fig. 1 shows the modified structure of the one based on the GIC analogue configuration, Antoniou and Rezk\(^1\) have proposed five second-order digital structures which realise lowpass, highpass, bandpass, notch and allpass transfer functions. It is known\(^2\) that all these five structures are free from zero-input limit cycles and only the highpass section is free from both zero and constant-input limit cycles.

We show now that lowpass and bandpass structures of Reference 2 can also be stabilised against constant-input limit cycles by making some simple modifications. The modified lowpass and bandpass structures are shown in Figs. 2 and 3, respectively, and their $A$, $B$ and $P$ matrices are given in eqns. 4 and 5, respectively:

$$A = \begin{bmatrix} -m_1 & -m_2 & 1 \\ 1 & m_1 & m_2 \end{bmatrix} \quad B = \begin{bmatrix} -1 & m_1 \\ 1 & 1 \end{bmatrix} \quad P = \begin{bmatrix} -1 \\ 0 \end{bmatrix}$$

$$A = \begin{bmatrix} -m_1 & -m_2 & 1 \\ 1 & m_1 & m_2 \end{bmatrix} \quad B = \begin{bmatrix} 1 & m_1 \\ 1 & 1 \end{bmatrix} \quad P = \begin{bmatrix} 1 \\ 0 \end{bmatrix}$$

The transfer functions $H_{low}(z)$ of Fig. 2 and $H_{band}(z)$ of Fig. 3 are as follows:

$$H_{low}(z) = \frac{-1}{z^2 + (m_1 - m_3)z + (1 + m_1 + m_2)}$$

$$H_{band}(z) = \frac{1}{z^2 + (m_1 - m_3)z + (1 + m_1 + m_2)}$$

Based on passive gyrator configurations, Kwan\(^3\) has proposed several digital biquads which are free of zero-input limit cycles. In fact the structure of Reference 1 is also one of these. These structures can also be modified in a similar way to make them free from constant-input limit cycles.

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References


ON ARRIVAL OF PATHS IN FADING MULTIPATH INDOOR RADIO CHANNELS

Indexing terms: Radiowave propagation, Radiocommunication, Radio links, Data transmission

Arrival of the paths in fading multipath channels obtained from several manufacturing floors and college campus laboratories at 910 MHz are studied. The discrepancies between the empirical distribution of the arriving paths and Poisson arrivals are discovered. The modified Poisson process is shown to fit the arriving paths closely.

Introduction: Recently, wideband measurements\(^1\)\(^-\)\(^3\) and statistical modelling\(^5\) of indoor radio channels has attracted tremendous attention for applications in offices, manufacturing
floors, warehouses, campuses and hospitals. All the reported wideband measurements and modelling are concerned with the statistics of the multipath spread. The only attempt in modelling the arrival of the paths in a building, is reported by Saleh and Valenzuela.

Saleh and Valenzuela use the mathematical model originally suggested by Turin for urban radio data communications. In this model, the complex envelope of the channel impulse response is represented by

$$h(t) = \sum_{k=1}^{L} a_k 6(t - \tau_k) e^{j\theta_k}$$

The transmitted impulse $\delta(t)$ is received as the sum of $L$ paths. The path $k$ has amplitude $a_k$, delay $\tau_k$ and carrier phase shift $\theta_k$. Based on the measurements in a research laboratory, the arrival of the paths is modelled as a Poisson process. The paths are presumed to have independent uniform phases, and independent Rayleigh amplitudes.

This letter reports the discrepancies following the Poisson arrival presumption and, based on empirical data collected from different manufacturing floors and college campus areas, presents a more accurate modified Poisson process to fit the arrival of the paths.

**Measurements:** The measurement set-up for the multipath propagation experiments, involved modulation of a 910 MHz signal by a train of 3 ns (3 dB width) pulses with 500 ns repetition period. The stationary receiver included a digital storage scope and a personal computer. The transmitter and the receiver used vertically polarised quarter-wave dipole antennas placed about 1.5m above floor level. The transmitter was moved to various locations in the site, and the received multipath profiles were stored in the computer. The distance between the transmitter and the receiver varied between 1 and 50m. A total of about 480 profiles were collected from measurements made in five areas on three different manufacturing floors and two areas on college laboratory floors. Each profile was an average over time of 64 profiles collected over 15-20s. A typical multipath profile, obtained from one of the locations is shown in Fig. 1.

$$\infty \int_{-\infty}^{\infty} |h(t)|^2 dt = \sum_{k=1}^{L} |a_k|^2$$

**Results and discussion:** For the path arrival distribution, the arrival time of the paths in each profile is divided into 5ns bins. The empirical curves were obtained by counting the number of paths in the first $N$ bins of each measured profile. The probability of occupancy in each bin was then plotted against the arriving delay (path index) and these points were connected by curves for clarity. The procedure is repeated for arriving paths in the first 5, 10, 15 and 20 bins. After careful visual inspection of many profiles, the threshold level employed to detect a genuine path in any bin was fixed at 30dB below the highest peak in the profile.

$$P_n(K) = \frac{\lambda^K}{K!} e^{-\lambda}$$

This is plotted as a continuous curve though it has values for only integer path numbers. We observe considerable discrepancies between the empirical and the Poisson distributions for all values of $N$. This discrepancy reflects a tendency of the paths to arrive in clusters, rather than in a random manner.

To explain similar discrepancies a modified Poisson model was proposed by Suzuki for urban radio channel modelling. According to the modified Poisson model, whenever there is a path in a bin, the mean arrival rate for the next bin is changed by a factor of $K$. The process becomes a standard Poisson sequence for $K = 1$. Depending on whether $K$ is greater or less than 1, the probability that there will be another path within the next bin increases or decreases, respectively. The underlying path occupancy rate $\lambda_n$ is calcu-
lated from the empirical path occupancy rate \( r_i \) by the following relation:

\[ \lambda_i = \frac{r_i}{(K - 1)r_{i-1} + 1}, \quad n \neq 1 \]

where \( \lambda_i = r_i \). The path number distribution is then calculated successively, using the recursive formulas given for this model in Reference 6. Figs. 4 and 5 show a comparison between the empirical path number distributions and the modified Poisson distributions for the manufacturing floors and college campuses, respectively. The curve fittings show considerable improvement over those of the Poisson model.

**Table 1** OPTIMUM VALUES OF \( K \)

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<tr>
<th>Area</th>
<th>Number of locations</th>
<th>Number of bins ( N )</th>
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<tr>
<td>Manufacturing floors</td>
<td>288</td>
<td>10</td>
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<td></td>
<td>15</td>
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**Conclusions:** Fading multipath profiles obtained from several different manufacturing floors and some college laboratories at 910 MHz, have been used to model the arrival of the paths. The empirical distribution was compared with the Poisson and the modified Poisson distributions and the modified at 910 MHz, have been used to model the arrival of the paths. The path number distribution is then calculated successively, using the recursive formulas given for this model in Reference 6. Figs. 4 and 5 show a comparison between the empirical path number distributions and the modified Poisson distributions for the manufacturing floors and college campuses, respectively. The curve fittings show considerable improvement over those of the Poisson model.

**References**


**Fig. 5** Comparison of empirical and modified Poisson distribution for college laboratory areas

This improvement is because the modified Poisson model utilizes the empirical probability of occupancy for each bin, whereas the Poisson model just uses the sum of the probabilities of occupancy for all bins. The optimum values of \( K \) calculated for the manufacturing floors and the campus environments are shown in Table 1.

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**Acknowledgment:** This work was supported in part by the National Science Foundation under contract NCR-8703435.

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**10MHz CMOS OTA-C VOLTAGE-CONTROLLED QUADRATURE OSCILLATOR**

**Indexing terms:** Circuit theory and design, Oscillators, Integrated circuits, Nonlinear networks

A quadrature-type voltage-controlled oscillator with operational transconductance amplifiers and capacitors (OTA-C) is presented. A monolithic integrated CMOS test circuit is introduced to verify theoretical results. The attainable frequency range of oscillation of the chip test circuit is 3-10.34 MHz. The total harmonic distortion (THD) is 0.20-1.87% for corresponding peak-to-peak amplitude voltages between 100 mV and 1 V. This amplitude can be controlled either by using a diode connection of two MOS transistors or a proposed nonlinear resistor.

**Introduction:** The quadrature mode of operation is a well-known mode of waveform generation in the field of sinewave oscillators. Two sinewaves in quadrature, that is with 90° phase difference, are generated. This is accomplished with a two-integrator loop. Besides this, some form of regeneration is usually included to ensure that the oscillation is created. The poles are initially located in the right half-plane (RHP) of the complex frequency plane and then pulled back by a nonlinear amplitude limiter.

To make a quadrature oscillator into a voltage-controlled oscillator (VCO), the integrators can be implemented using OTAs. The output current of a CMOS-OTA is controlled by the differential input voltage and its transconductance gain \( g_m \). Furthermore, \( g_m \) can be varied over several octaves by adjusting an external DC amplifier bias current \( I_{bias} \), i.e.

\[ g_m = h_c \sqrt{I_{bias}} \]  

(1)

where \( h_c \) is a process-dependent parameter.

In this letter an integrated CMOS OTA-C quadrature VCO using a 3-µm double metal technology (processed by MOSIS, Marina del Rey, CA) is presented. The frequency of oscillation can be externally adjusted by nearly two octaves with a THD less than 1.9%. The real part of the poles can be adjusted, in practice, independently of the frequency of oscillation.

**Fig. 1** Proposed OTA-C quadrature oscillator structure

**Theory:** The proposed quadrature VCO architecture is shown in Fig. 1. It has one inverting \( g_{m1}, C_1 \) and one nonlinear integrating \( g_{m2}, C_2 \) and one noninverting \( g_{m3}, C_3 \). \( R_n \) is a nonlinear resistor whose \( i/V \) characteristics are depicted in Fig. 2a. \( R_n \) provides the nonlinear amplitude limiter function. Implementations are discussed later. The OTAs associated with \( g_{m2} \) and \( g_{m4} \) allow