Cascadable current-mode filters using single FTFN

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Indexing terms: Current-mode circuits, Filters

A new configuration for realising current-mode filters using a single four-terminal floating nullor (FTFN) is presented. It can realise lowpass, bandpass, highpass, notch and allpass filters from the same configuration. This configuration has a high output impedance, so the synthesised current-mode filters can be cascaded without additional buffers. Moreover, the resultant current-mode filters will be insensitive to the current tracking error of an FTFN. Experimental results that confirm the theoretical analyses are obtained.

Introduction: Current-mode circuits have been receiving significant attention as they have the potential advantages of accuracy and wide bandwidth over their voltage-mode counterparts [1]. Potentially, there is one limitation in most current-mode circuits. They need additional current buffers to cascade the similar stages or to extract the output current. Therefore, several cascadable current-mode filters [2 – 6] using second-generation current conveyors (CCIIs) and a single four-terminal floating nullor (FTFN) [7 – 9] have been developed. However, there is no configuration using a single FTFN, which can realise various types of cascadable current-mode filters. In this Letter, such a new configuration is presented. Experimental results which confirm the theoretical analysis are obtained.

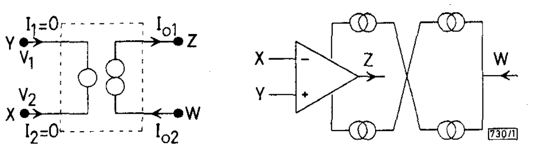

Fig. 1 *Nullor model of FTFN*

Circuit description: The port relations of an FTFN, shown in Fig. 1, can be characterised as $I_1 = I_2 = 0$, $V_1 = V_2$ and $I_{o1} = I_{o2}$ [7 – 9]. Considering the proposed configuration in Fig. 2, its transfer function can be derived as

$$\frac{I_o}{I_{in}} = \frac{y_5 + y_2(\frac{y_5}{y_1} - \frac{y_4}{y_3})}{y_4 + y_2(\frac{y_4}{y_3} - \frac{y_5}{y_1}) + y_6(1 + \frac{y_4}{y_3})} \qquad (1)$$

If $y_6 = 0$ (open circuit) and $y_4/y_3 - y_5/y_1$ (i.e. $z_3 = z_4(1 + z_1/z_5)$), eqn. 1 can be simplified as

$$\frac{I_o}{I_{in}} = \frac{y_5 - y_2}{y_4 + y_2} \qquad (2)$$

A first-order allpass filter can be obtained if $y_1 = 1/R_1$, $y_2 = sC$, $y_4 = y_5 = 1/R$ and $y_3 = 1/(R_1 + R)$.

If $y_6 = 0$, $y_1 = y_2$ and $y_4 = y_5$, eqn. 1 can be simplified as

$$\frac{I_o}{I_{in}} = \frac{2y_3 - y_2}{y_2} \qquad (3)$$

Furthermore, if $y_2 = sC_2 + 1/R_2$ and $y_3 = 1/(R_3 + 1/sC_3)$, eqn. 3 will be

$$\frac{I_o}{I_{in}} = -\frac{s^2C_2C_3R_2R_3 + s(C_2R_2 + C_3R_3 - 2C_3R_2) + 1}{s^2C_2C_3R_2R_3 + s(C_2R_2 + C_3R_3) + 1} \qquad (4)$$

Hence, if $C_2R_2 + C_3R_3 = 2C_3R_2$, a second-order notch filter can be realised. If $C_2R_2 + C_3R_3 = C_3R_2$, a second-order allpass filter can be achieved.

If $y_6 = 0$ and $y_4/y_3 = y_5/y_2$, eqn. 1 can be simplified as

$$\frac{I_o}{I_{in}} = \frac{y_5y_2}{y_4y_1 + y_5y_1 - y_2y_5} \qquad (5)$$

Furthermore, if $y_1 = sC_1 + 1/R_1$, $y_2 = sC_2$, $y_3 = 1/R_3$, $y_4 = 1/R_4$, $y_5 = sC_5$ and $C_2R_3 = C_5R_4$, eqn. 5 can be expressed as

$$\frac{I_o}{I_{in}} = \frac{s^2C_2C_5R_1R_4}{s^2(C_1 - C_2)C_5R_1R_4 + s(C_1R_1 + C_5R_4) + 1} \qquad (6)$$

Thus, if $C_1 > C_2$, a second-order highpass filter can be obtained.

If $y_1 = sC_1 + 1/R_1$, $y_2 = 1/R_2$, $y_3 = sC_3$, $y_4 = sC_4$, $y_5 = 1/R_5$ and $C_4R_5 = C_3R_2$, eqn. 5 can be expressed as

$$\frac{I_o}{I_{in}} = \frac{1}{s^2C_1C_4R_2R_5 + s(C_4\frac{R_2R_5}{R_1} + C_1R_2) + (\frac{R_2}{R_1} - 1)} \qquad (7)$$

Thus, if $R_2 > R_1$, a second-order lowpass filter can be obtained.

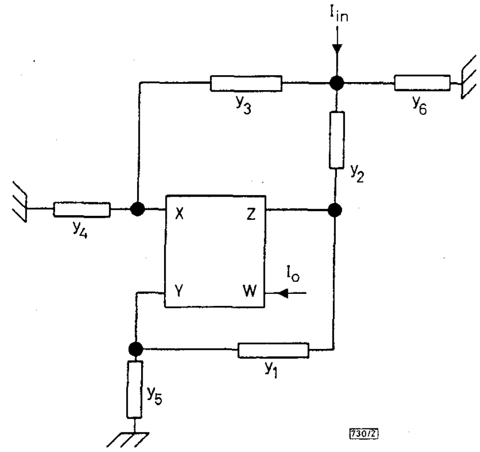

Fig. 2 *Configuration for realising current-mode filters with high output impedance*

To realise a bandpass filter, the admittances might be chosen as follows: $y_1 = sC_1 + 1/R_1$, $y_2 = y_3 = y_6 = 1/R_2$, and $y_4 = y_5 = sC_4$. eqn. 1 can be expressed as

$$\frac{I_o}{I_{in}} = \frac{sC_4/R_2}{s^23C_1C_4 + s(\frac{3C_4}{R_1} + \frac{C_1}{R_2} - \frac{C_4}{R_2}) + \frac{1}{R_1R_2}} \qquad (8)$$

The natural frequency and quality factor of this bandpass filter can be expressed as

$$\omega_o = \frac{1}{\sqrt{3C_1C_4R_1R_2}} \quad \text{and} \quad Q = \frac{\sqrt{3C_1C_4R_1R_2}}{3C_4R_2 + C_1R_1 - C_4R_1}$$

Taking into account the nonideal FTFN, namely $I_{o2} = \alpha I_{o1}$ and $V_2 = \beta V_1$, where $\alpha = 1 - \varepsilon_1$, ε_1 ($\varepsilon_1 << 1$) denotes the current tracking error of an FTFN and $\beta = 1 - \varepsilon_2$, ε_2 ($\varepsilon_2 << 1$) is the input voltage tracking error, since the current output of the FTFN in Fig. 1 is directly connected to the load, the resultant current-mode filters will be insensitive to the current tracking error of an FTFN except for current gain.

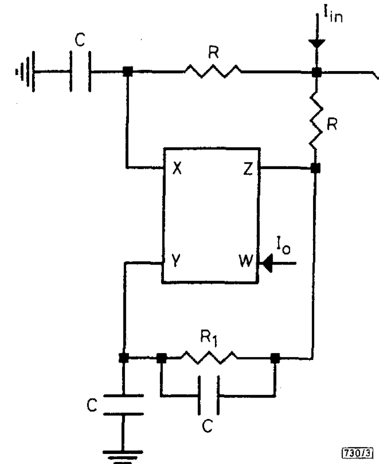

Fig. 3 *Current-mode second-order bandpass filter*

Experimental results: To demonstrate the feasibility of the proposed circuits, a current-mode bandpass filter derived from Fig. 1 is implemented as shown in Fig. 3 with $C = 1$nF, $R = 10$ kΩ and $R_1 = 270$kΩ. The FTFN consists of an operational amplifier (LF356) and bipolar transistors (CA3096AE) [6, 9]. Fig. 4 shows the experimental results for a second-order bandpass filter in Fig. 3. Experimental results confirm with the results of the theoretical analysis.

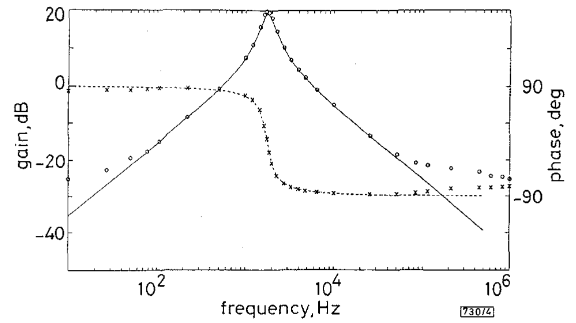

Fig. 4 *Comparisons between theoretical and experimental results for bandpass filter in Fig. 3 with C = 1nF, R = 10kΩ and R_I = 270kΩ*

Theory: ——— gain, – – – – phase
Experiment: ○ gain, × phase

Conclusions: A new configuration with single FTFN for realising various types of cascadable current-mode filters is presented. Second-order lowpass, bandpass, highpass, allpass and notch filters can be achieved from the same configuration. Experimental results confirm the results of the theoretical analysis. The proposed circuits are expected to be useful in analogue filtering applications.

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Analysis of an arbitrarily inclined slot on the ground plane of a microstrip line

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Indexing terms: Slot antennas, Spectral-domain analysis

The analysis of an arbitrarily inclined slot on the ground plane of a microstrip line is presented using the spectral domain approach and reciprocity theorem. Comparison of numerical results for the special case of a perpendicular slot with other available computed and measured data shows a good agreement. It is found that the impedance level can be controlled over a wide range by varying inclination angle. Results should be useful in antenna and multilayer circuit integration applications.

Introduction: In recent years, printed slots on the ground plane of a microstrip line have been shown to be very useful in many applications. For instance, slots can be used as radiating elements in certain planar arrays, to excite microstrip patch and dielectric resonator antennas and to couple the energy between multilayer integrated circuits. Several rigorous numerical and approximate methods [1 – 5] have been applied to analyse microstrip line fed printed slots, and results for a centre-fed slot perpendicular to a microstrip line have shown very high resonant radiation resistance. To realise the efficient impedance matching and to control the coupling, the slot may need to be inclined with a <90° angle to the microstrip line. However, most analyses reported so far are based on the assumption that the slot is perpendicular to the microstrip line, and few limited results were reported for an arbitrarily inclined slot [6].

The purpose of this Letter is to present a new general analysis of an arbitrarily inclined slot on the ground plane of a microstrip line. The spectral domain approach is used to obtain propagation characteristics of the fundamental mode of the microstrip line and the fields generated by the slot. The reciprocity theorem is applied to derive the reflection and transmission coefficients, from which the components of an equivalent slot circuit are found.

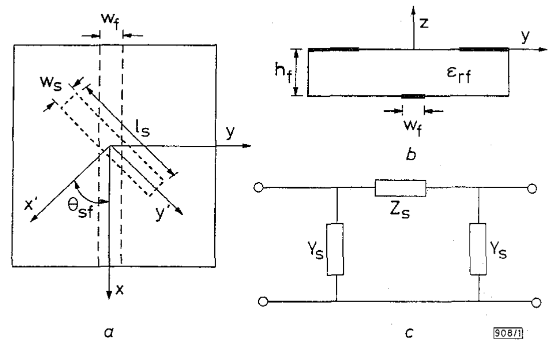

Fig. 1 *Top view and end view of an inclined slot in the ground plane of an infinite microstrip line, and the equivalent circuit of the slot*

a Top view
b End view
c Equivalent circuit

Analysis: The geometry of an inclined slot (inclination angle = 90° – θ_{sf}) in the ground plane of a microstrip line is shown in Fig. 1*a* and *b*. The microstrip line is assumed to be infinitely long. The spectral domain approach is first used to obtain the propagation constant β, characteristic impedance $Z_c = 1/Y_c$ and Fourier-transformed magnetic field components $\tilde{h}(\kappa_y, z)$, $\tilde{h}_y(\kappa_y, z)$ of the fundamental (quasi-TEM) mode of the microstrip line propagating in the positive *x* direction. In the Galerkin procedure in the spectral domain for the solution of the above parameters, the electric current components on the line are expanded in terms of Chebyshev polynomials weighted by an appropriate singular function. The slot length is l_s, and the slot width w_s is assumed here to be electrically small, hence only the *x'* component of the electric field $E^s_{x'}$ in the slot will be considered and expanded in terms of piecewise sinusoidal functions. By using reciprocity [1], we can obtain the reflection and transmission coefficients at the slot:

$$R = \frac{1}{4\pi}\int_{-\infty}^{\infty}\frac{\sin F_1}{F_1}\tilde{E}^s_{x'}(-\kappa_y\cos\theta_{sf} - \beta\sin\theta_{sf})F_2 d\kappa_y \quad (1)$$

$$T = 1 - \frac{1}{4\pi}\int_{-\infty}^{\infty}\frac{\sin F_1}{F_1}\tilde{E}^s_{x'}(\kappa_y\cos\theta_{sf} + \beta\sin\theta_{sf})F_2 d\kappa_y \quad (2)$$

where

$$F_1 = 0.5w_s(-\kappa_y\sin\theta_{sf} + \beta\cos\theta_{sf}) \quad (3)$$

$$\tilde{E}^s_{x'}(\alpha) = \int_{-0.5l_s}^{0.5l_s} E^s_{x'}(y')e^{j\alpha y'}dy' \quad (4)$$

$$F_2 = \cos\theta_{sf}\tilde{h}_y(\kappa_y, 0) + \sin\theta_{sf}\tilde{h}_x(\kappa_y, 0) \quad (5)$$

The equivalent circuit of the slot can take the form of a symmetrical π circuit, as shown in Fig. 1*c*. Then, the series impedance Z_s and shunt admittance Y_s of this circuit can be obtained from *R* and *T*: