flowchart is similar to that given in Fig.3.

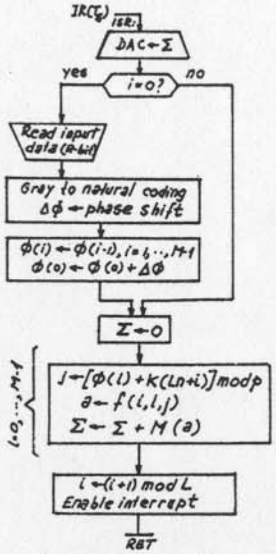

Fig.3. The cyclical processing program segment (PSK)

6. CONCLUSIONS

The data signal generation from stored elements avoids multiplications and permit a high processing speed.

The memory size is small, with a proper choice of f_s/f_b ratio. Hence, the method is useful in implementation of modems.

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Author: Asst. E.Borcoci, Address: Polytechnical Institute of Bucharest, Electronic Dept., Splaiul Independentei 313, 77748, Bucharest, ROMANIA.

J. Vandewalle, H. De Man and J. Rabaey

ESAT LABORATORY, Katholieke Universiteit Leuven, Belgium.

A UNIFIED THEORY FOR THE COMPUTER AIDED ANALYSIS OF GENERAL SWITCHED CAPACITOR NETWORKS

ABSTRACT

We derive a general theory for the analysis of switched capacitor circuits. The circuits can have 2 phases or more and have no topologic constraints whatsoever. Also the inputs are arbitrary. We describe efficient computations of the time domain response, frequency response, noise and sensitivity analysis. All derivations are carried out in the modified nodal analysis framework, which allows an easy implementation in a computer program. The theory is simple and implies all previously described analysis techniques.

1. INTRODUCTION

Switched capacitor (SC) circuits have received much attention in recent years, because they allow to realize a filter with low sensitivity as an integrated circuit. Since these circuits are usually large, the analysis should be done by computer. Two types of computer implementations have been described [1-5], one based on time-domain analysis and one based on z-domain analysis. Here we present a unifying framework for both, which then also allows to evaluate their relative merits.

2. THEORETICAL ASPECTS OF THE ANALYSIS OF SC-NETWORKS

We consider arbitrary linear networks containing ideal switches, capacitors, independent voltage and charge sources and dependent sources VCVS, QCQS, ... The switches are controlled by Boolean clock variables $\phi_r(t) = 0$ or 1. $\phi_r(t) = 0$ (resp., $\phi_r(t) = 1$) corresponds to an open (resp., closed) switch at time t if this switch is driven by clock r. The time is partitioned into time slots $\Delta_k = (t_k, t_{k+1}]$ such that the clock signals do not vary in Δ_k, i.e. $\phi_r(t) = \phi_{rk}$ for $t \in \Delta_k$. We assume that the clock signals are T-periodic, with N time slots (called N phases) in one period of duration T.

Theorem 1 : If any above defined SC network is excited by a piecewise-constant excitation ($\underline{v}(t) = \underline{v}_m$, $\underline{u}(t) = \underline{u}_m$, $t \in \Delta_m$), then its response in the time domain is also piecewise-constant $\mathbf{v}(t) = \mathbf{v}_m$, $\mathbf{q}(t) = \mathbf{q}_m$ in Δ_m and given by

$$\left[\begin{array}{c|c} \mathbf{A}_k & \mathbf{B}_k \\ \hline \mathbf{C}_k & \mathbf{D}_k \end{array}\right] \left[\begin{array}{c} \mathbf{v}_{k+\ell N} \\ \hline \mathbf{q}_{k+\ell N} \end{array}\right] = \left[\begin{array}{c} \mathbf{A}_k \ \mathbf{v}_{k+\ell N-1} \\ \hline \mathbf{0} \end{array}\right] + \left[\begin{array}{c} \mathbf{w}_{k+\ell N} \\ \hline \mathbf{u}_{k+\ell N} \end{array}\right] \qquad (1)$$

where $\mathbf{v}(t)$ (resp., $\mathbf{u}(t)$) is the vector of the voltage responses at the nodes (resp., voltage sources in some selected branches) and where $\mathbf{q}(t)$ (resp., $\mathbf{w}(t)$) is the vector of the charges transferred in some selected branches (resp., injected by charge sources in the nodes) between $t_{k+\ell N}$ and t for t $\varepsilon\ \Delta_{k+\ell N}$, and where the contributions to $\mathbf{A}_k$, $\mathbf{B}_k$, $\mathbf{C}_k$ and $\mathbf{D}_k$ are given by the stamps of the classical modified nodal analysis and where the contribution of a switch S connecting node i to j and governed by clock ϕ_r, is given by

$$\begin{array}{c} \\ j \\ j \\ S \end{array} \begin{array}{c} \begin{array}{ccc} i & j & S \end{array} \\ \left[\begin{array}{cc|c} 0 & 0 & \phi_{rk} \\ 0 & 0 & -\phi_{rk} \\ \hline \phi_{rk} & -\phi_{rk} & \bar{\phi}_{rk} \end{array}\right] \end{array} \left[\begin{array}{c} v_{ik} \\ v_{jk} \\ \hline q_{Sk} \end{array}\right] = \left[\begin{array}{c} 0 \\ 0 \\ \hline 0 \end{array}\right] \qquad (2)$$

where ‾ denotes the complement of a Boolean variable.
If the SC network is excited by an arbitrary excitation $\mathbf{w}(t)$, $\mathbf{u}(t)$, then the piecewise-constant part of the input and output $\mathbf{w}_m = \mathbf{w}(t^-_{m+1})$, $\mathbf{u}_m = \mathbf{u}(t^-_{m+1})$, $\mathbf{v}_m = \mathbf{v}(t^-_{m+1})$ and $\mathbf{q}_m = \mathbf{q}(t^-_{m+1})$ are determined by (1) and the remainder wave forms $\mathbf{w}^*(t) = \mathbf{w}(t)-\mathbf{w}_m$, $\mathbf{u}^*(t) = \mathbf{u}(t)-\mathbf{u}_m$, $\mathbf{v}^*(t) = \mathbf{v}(t)-\mathbf{v}_m$, and $\mathbf{q}^*(t) = \mathbf{q}(t)-\mathbf{q}_m$ for t $\varepsilon\ \Delta_m$ are related by

$$\left[\begin{array}{c|c} \mathbf{A}_k & \mathbf{B}_k \\ \hline \mathbf{C}_k & \mathbf{D}_k \end{array}\right] \left[\begin{array}{c} \mathbf{v}^*(t) \\ \hline \mathbf{q}^*(t) \end{array}\right] = \left[\begin{array}{c} \mathbf{w}^*(t) \\ \hline \mathbf{u}^*(t) \end{array}\right] \quad \text{for } t \ \varepsilon \ \Delta_{k+\ell N} \qquad (3)$$

Proof : The first set of equations of (1,3) express the charge conservation and the second set are the constitutive relations of some selected branches. □

Theorem 2 : Given any above defined SC network with a piecewise-constant excitation. Define

$$\mathbf{V}_m(z) \triangleq \mathcal{Z}\{\mathbf{v}_{m+\ell N}\} = \sum_{\ell=0}^{\infty} \mathbf{v}_{m+\ell N} \ z^{-\ell} \qquad (4)$$

and analogous equations for $\mathbf{U}_m$, $\mathbf{Q}_m$ and $\mathbf{W}_m$ where m = 1, 2, ... N. Then these variables are related by :

$$\left[\begin{array}{ccccc|cccc} \mathbf{A}_1 & & & & -\mathbf{A}_1 z^{-1} & \mathbf{B}_1 & & & \\ -\mathbf{A}_2 & \mathbf{A}_2 & & & & & \mathbf{B}_2 & & \\ & -\mathbf{A}_3 & \mathbf{A}_3 & & & & & \mathbf{B}_3 & \\ & & \ddots & \ddots & & & & & \ddots \\ & & & -\mathbf{A}_N & \mathbf{A}_N & & & & \mathbf{B}_N \\ \hline \mathbf{C}_1 & & & & & \mathbf{D}_1 & & & \\ & \mathbf{C}_2 & & & & & \mathbf{D}_2 & & \\ & & \mathbf{C}_3 & & & & & \mathbf{D}_3 & \\ & & & \ddots & \mathbf{C}_N & & & \ddots & \mathbf{D}_N \end{array}\right] \left[\begin{array}{c} \mathbf{V}_1 \\ \mathbf{V}_2 \\ \mathbf{V}_3 \\ \vdots \\ \mathbf{V}_N \\ \hline \mathbf{Q}_1 \\ \mathbf{Q}_2 \\ \mathbf{Q}_3 \\ \vdots \\ \mathbf{Q}_N \end{array}\right] = \left[\begin{array}{c} \mathbf{W}_1 \\ \mathbf{W}_2 \\ \mathbf{W}_3 \\ \vdots \\ \mathbf{W}_N \\ \hline \mathbf{U}_1 \\ \mathbf{U}_2 \\ \mathbf{U}_3 \\ \vdots \\ \mathbf{U}_N \end{array}\right] \qquad (5)$$

where the missing entries are zero, and the submatrices are defined in Theorem 1.

Proof : Apply the techniques of z-transform and (4) on (1). □

Corollary 1 : Any above defined SC-network is completely characterized by the z-domain transfer matrix $\mathbf{M}$ of

$$\left[\begin{array}{c} \mathbf{V}_1 \\ \mathbf{V}_2 \\ \vdots \\ \mathbf{V}_N \\ \hline \mathbf{Q}_1 \\ \mathbf{Q}_2 \\ \vdots \\ \mathbf{Q}_N \end{array}\right] = \underbrace{\left[\begin{array}{cccc|cccc} \mathbf{G}_{11} & \mathbf{G}_{12} & \cdots & \mathbf{G}_{1N} & \mathbf{H}_{11} & \mathbf{H}_{12} & \cdots & \mathbf{H}_{1N} \\ \mathbf{G}_{21} & \mathbf{G}_{22} & \cdots & \mathbf{G}_{2N} & \mathbf{H}_{21} & \mathbf{H}_{22} & \cdots & \mathbf{H}_{2N} \\ \cdots & \cdots & \cdots & \cdots & \cdots & \cdots & \cdots & \cdots \\ \mathbf{G}_{N1} & \mathbf{G}_{N2} & \cdots & \mathbf{G}_{NN} & \mathbf{H}_{N1} & \mathbf{H}_{N2} & \cdots & \mathbf{H}_{NN} \\ \hline \mathbf{K}_{11} & \mathbf{K}_{12} & \cdots & \mathbf{K}_{1N} & \mathbf{L}_{11} & \mathbf{L}_{12} & \cdots & \mathbf{L}_{1N} \\ \mathbf{K}_{21} & \mathbf{K}_{22} & \cdots & \mathbf{K}_{2N} & \mathbf{L}_{21} & \mathbf{L}_{22} & \cdots & \mathbf{L}_{2N} \\ \cdots & \cdots & \cdots & \cdots & \cdots & \cdots & \cdots & \cdots \\ \mathbf{K}_{N1} & \mathbf{K}_{N2} & \cdots & \mathbf{K}_{NN} & \mathbf{L}_{N1} & \mathbf{L}_{N2} & \cdots & \mathbf{L}_{NN} \end{array}\right]}_{\mathbf{M}} \left[\begin{array}{c} \mathbf{W}_1 \\ \mathbf{W}_2 \\ \vdots \\ \mathbf{W}_N \\ \hline \mathbf{U}_1 \\ \mathbf{U}_2 \\ \vdots \\ \mathbf{U}_N \end{array}\right] \qquad (6)$$

Proof : Eq. (6) is obtained by inverting the matrix in (5). Eq. (3) can easily be recovered from $\mathbf{M}$ at ∞. □

The submatrices $\mathbf{G}_{k\ell}$, $\mathbf{H}_{k\ell}$, $\mathbf{K}_{k\ell}$ and $\mathbf{L}_{k\ell}$ have a very simple interpretation. If only nonzero voltage sources are supplied in time slots ℓ, ℓ+N, ℓ+2N, ... and if the output node voltages are only observed at the end of time slots k, k+N, k+2N, ... then $\mathbf{H}_{k\ell}(z)$ relates inputs at phase ℓ to outputs at phase k i.e.

$$\underset{\sim}{V}_k(z) = \underset{\sim}{H}_{k\ell}(z)\, \underset{\sim}{U}_\ell(z) \qquad (7)$$

It is shown in [6] that for any SC network $\mathcal{N}$ and adjoint SC network $\tilde{\mathcal{N}}$ can be derived with a z-domain transfer matrix $\tilde{\underset{\sim}{M}}$, whose columns correspond with rows of $\underset{\sim}{M}$.

3. COMPUTATION OF FREQUENCY, NOISE AND SENSITIVITY PROPERTIES OF SC NETWORKS

Since a SC network is a time-varying circuit, a sinusoidal excitation may generate many frequencies in the output. However often one is only interested in the following practical frequency domain transfer function $\underset{\sim}{H}(\omega)$ [1,3], which relates by definition the phasor $\underset{\sim}{U}$ of the sinusoidal excitation $\underset{\sim}{u}(t) = \underset{\sim}{U}\, e^{j\omega t}$ to the phasor at the same pulsation ω in the output $\underset{\sim}{v}(t)$. Standard Fourier transform techniques allow to prove that

$$\underset{\sim}{H}(\omega) = \sum_{k=1}^{N} e^{-j\omega t_{k+1}}\, \nu_k(\omega) \left(\sum_{\ell=1}^{N} e^{j\omega t_{\ell+1}}\, \underset{\sim}{H}_{k\ell}(e^{j\omega T})\right) + [(t_{k+1}-t_k)/T - \nu_k(\omega)]\underset{\sim}{H}_{kk}(\infty), \qquad (8.a)$$

where

$$\nu_k(\rho) = 2\{\sin[\rho(t_{k+1}-t_k)/2]\ \exp[j\rho(t_{k+1}-t_k)/2]\}/T\rho. \qquad (8.b)$$

Similar expressions can be derived for the aliasing terms. In the computation of one entry (i,j) of (8.a) via the time-domain, N impulse responses have to be determined using (1) for impulse excitations in the N phases. The $H^{(ij)}_{k\ell}$ of (8.a) are then obtained via an FFT. In the z-domain we solve (5) once for $z = e^{j\omega T}$ and $\underset{\sim}{W}_\ell = \underset{\sim}{0}$, $U^{(n)}_\ell = 0$ for $n \neq j$ and $U^{(j)}_\ell = e^{j\omega t_{\ell+1}}$ for $\ell = 1, \ldots N$ in order to obtain the sums in brackets in (8.a).

In noise analysis one is usually interested in the output noise voltage at one node i due to the noise in many (say s) components. Let the vector $\underset{\sim}{u}(t)$ of the input noise voltage sources be characterized by a given spectral density matrix $\underset{\sim}{S}_{\underset{\sim}{u}}(\omega)$ which depends on the technology. Then the average spectral density function is given by

$$S_{v^{(i)}}(\omega) = \sum_{n=-\infty}^{+\infty} \underset{\sim}{X}^{(i.)}(\omega, \omega - n\tfrac{2\pi}{T})\ \underset{\sim}{S}_{\underset{\sim}{u}}(\omega - n\tfrac{2\pi}{T})\ \underset{\sim}{X}^{(i.)*}(\omega, \omega - n\tfrac{2\pi}{T}) \qquad (9.a)$$

where

$$\underset{\sim}{X}^{(i.)}(\omega,\Omega) = \sum_{\ell=1}^{N} e^{j\Omega t_{\ell+1}} \left(\sum_{k=1}^{N} \nu_k(\omega)\, e^{-j\omega t_{k+1}}\, \tilde{\underset{\sim}{K}}^{(.i)T}_{\tilde{\ell}\tilde{k}}(e^{j\Omega T})\right) + \sum_{k=1}^{N} [\nu_k(\omega-\Omega) - \nu_k(\omega)]\, e^{j(\Omega-\omega)t_{k+1}}\, \tilde{\underset{\sim}{K}}^{(.i)}_{kk}{}^{T}(\infty) \qquad (9.b)$$

and * denotes complex conjugate and transpose and $\tilde{\underset{\sim}{K}}_{\tilde{\ell}\tilde{k}}$ with $\tilde{\ell} = N-\ell+1$, $\tilde{k} = N-k+1$ is the corresponding submatrix of the z-domain transfer matrix of the adjoint SC network and $\underset{\sim}{A}^{(i.)}$ (resp., $\underset{\sim}{A}^{(.j)}$) is the i-th row (resp., j-th column) of the matrix $\underset{\sim}{A}$. In the time domain analysis all required $\tilde{\underset{\sim}{K}}^{(.i)}_{\tilde{\ell}\tilde{k}}$ can be obtained from N impulse responses of the adjoint SC network and FFT's. In the z-domain analysis all required sums in brackets in (9.b) are obtained by solving for each frequency one set of equations (5) for the adjoint SC network with an appropriate excitation.

In sensitivity analysis one is interested in the effect of a variation in one or several components or the frequency on the input output relation. They can be computed by first deriving the sensitivity of an arbitrary entry $H^{(ij)}_{k\ell}$ of the z-domain transfer function (6) and then using these equations to compute the sensitivity of the frequency domain transfer function. $H^{(ij)}(\omega) = A^{(ij)}(\omega)e^{j\phi^{(ij)}(\omega)}$. We illustrate this by computing the group delay

$$T^{(ij)}(\omega) \triangleq \frac{\partial \phi^{(ij)}(\omega)}{\partial \omega} = \mathrm{Im}\left(\frac{1}{H^{(ij)}(\omega)} \sum_{k=1}^{N} \frac{\partial H^{(ij)}_k(\omega)}{\partial \omega}\right) \qquad (10.a)$$

where

$$\frac{\partial H^{(ij)}_k(\omega)}{\partial \omega} = \frac{[\omega(t_{k+1}-t_k)+j]\, e^{j\omega(t_{k+1}-t_k)} - j}{T\omega^2} \left\{\sum_{\ell=1}^{N} e^{j\omega(t_{\ell+1}-t_{k+1})} H^{(ij)}_{k\ell}(e^{j\omega T}) - H^{(ij)}_{kk}(\infty)\right\}$$
$$+ j\,\nu_k(\omega)\left\{\sum_{\ell=1}^{N} e^{j\omega(t_{\ell+1}-t_{k+1})} \left[(t_{\ell+1}-t_{k+1})\, H^{(ij)}_{k\ell}(e^{j\omega T}) + T e^{j\omega T} \frac{\partial H^{(ij)}_{k\ell}(z)}{\partial z}\Big|_{z=e^{j\omega T}}\right]\right\} \qquad (10.b)$$

$$\frac{\partial H^{(ij)}_{k\ell}(z)}{\partial z} = -z^{-2}\, \tilde{\underset{\sim}{G}}^{(.i)\,T}_{1\tilde{k}}\ \underset{\sim}{A}_1\ \underset{\sim}{H}^{(.j)}_{N\ell} \qquad (10.c)$$

In order to obtain the group delay and all other sensitivities via the time domain, N impulse responses of the nominal and N impulse responses of the adjoint SC network and some FFT's have to be computed. In the z-domain analysis, (10.a) is obtained by solving three sets of equations (5) for the nominal SC network and one for the adjoint network.

4. CONCLUSION AND COMPARISON

We have described an elegant and general theory for the analysis of SC circuits from which the computer implementations via time domain or via z-domain are derived. Other analysis techniques [2-5] are either restricted to two phase SC networks [4-5] or require lengthy derivations [2-3] in order to obtain equations equivalent to (1) or (5). Although [2-3] may have less equations to solve, we are convinced that the use of macromodels in our framework can provide us with similar reductions. Moreover, the search for the minimal set of equations is largely academic, since sparse matrix methods have a computational complexity, which depends on the zero-nonzero pattern and not on the size of the matrix. It is important to observe that the use of the adjoint SC network [6] allowed great computational savings and hence it is equally useful in [2- 5]. Let us conclude by comparing the time and z-domain technique. The time domain requires repeated solutions of (1), which is a smaller set of equations than (5), but it requires additional FFT's. The time domain analysis is implemented in DIANA [1] and progress is being made toward an implementation of the z-domain analysis.

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ESAT Laboratory, K. Univ. Leuven

Kardinaal Mercierlaan 94

B-3030 Heverlee, Belgium

E. V. Sørensen
326, Institute of Circuit Theory and Telecommunication
Technical University of Denmark
DK-2800 Lyngby - Denmark

NORATOR-NULLATOR FORMULATION OF THE NETWORK FUNCTIONS FOR SWITCHED-CAPACITOR CIRCUITS

ABSTRACT: A norator-nullator approach is used to express the z-domain network functions of switched capacitor circuits as a signed ratio of two determinants, i.e. in a form which permits the determination of poles and zeros by means of the well known two-sets-of-eigenvalues-approach. The involved nodal capacitance matrices are completely reduced with respect to all constraints imposed by closed switches, operational amplifiers and port-conditions. An algorithm is described, whereby these matrices can be compiled directly from the input circuit description.

1. Introduction

Recently several approaches to computer-aided analysis of switched-capacitor circuits have appeared [4][5][6] and undoubtedly many more will show up. The method described here is aimed at z-domain analysis of sampled-input, two-phase, equal time-slot, switched-capacitor two-ports consisting of capacitors, ideal switches and ideal operational amplifiers. It is based on Kurth's and Moschytz's fundamental nodal approach [1], and is related to Hőkenek's and Moschytz's later computer-oriented method [5], but emphasizes and simplifies the algorithmic aspects in obtaining the necessary matrices. The theory is developed for two-phase switching schemes, but the extention to multiphase systems is obvious.

2. Universal formulas for switched-capacitor network functions

Fig. 1 shows the equivalent z-domain four-port [1] corresponding to a two-phase switched-capacitor two-port with equal time-slots A and B, and defines the port-node designations and sign-conventions adopted in the following. Except for the datum node: 0, every node k (k=1,2••n) in the physical two-port has an equivalent A-node: $k^A=k$ and an equivalent B-node: $k^B=k+n$ in the four-port.

Fig. 2 defines the transfer function $H^{BA}(z) = M^B(z)/G^A(z)$ with the generator and meter quantities: G^A and M^B being of type V or Q. Proper definition of H^{BA} requires that input-port B is short-circuited if G^A is a voltage, and that output port A is short-circuited if M^B is a charge. This is taken care of by the hypothetical switches SG and SM respectively. With these conditions satisfied H^{BA} may be found by nodal analysis of the equivalent two-port defined by ports p^A, q^A and r^B, s^B. Since similar definitions apply for H^{AA}, H^{AB} and H^{BB} we now drop the phase designation for H, pq and rs.

In order to obtain the reduced nodal z-domain capacitance matrix, let us first remove all closed switches and operational amplifiers from the equivalent two-port. This leaves an all-capacitive (possibly disconnected) structure with a 2n·2n nodal capacitance matrix $\underline{\underline{C}}(z)$ of the form shown in Eq. 1:

$$\underline{\underline{C}}(z) = \left[\begin{array}{c|c} \underline{\underline{C}}_0 & z^{-1/2}\cdot(-\underline{\underline{C}}_0) \\ \hline z^{-1/2}\cdot(-\underline{\underline{C}}_0) & \underline{\underline{C}}_0 \\ \underbrace{\qquad}_{A} & \underbrace{\qquad}_{B} \end{array}\right] \begin{array}{l} \}\,A \\ \}\,B \\ \end{array} \qquad (1)$$

where $\underline{\underline{C}}_0$ is the simple real n×n nodal capacitor matrix of the correspondingly stripped physical circuit.

Next, let us model the short-circuits and op-amps by means of nullator-norator pairs as shown in Fig. 3, and reconstruct the equivalent two-port by