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A QUASILOSSLESS TIME-VARIABLE SYNTHESIS SUITABLE FOR INTEGRATED CIRCUITS

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Abstract—

The essence of man is his heart
And what thereto pertains.

Here using capacitors, normalized fixed, and gyrators,
grounded time-varying, the two shaping hearts
integrated in linear forms, a practical Foster
type synthesis gains: quasilossless, n-port,
integratable, time-variable, minimal C.

Though hearts are practiced bare known
Integrated their essence obtains.

1. INTRODUCTION

THE importance of microsystems to today's technology, as well as the ability to realize scientific dreams, has led to an awareness of the vast potential of integrated circuits[1, 2], their popularization[3] and their interpretation in art.[4] Because of the complex logic size and reliability requirements of modern digital systems and the availability to fulfil many of these requirements [Ref. (5), pp. 190–203], integrated circuits have found, and should continue to find, wide acceptance in digital computer design.[6, 7] In the linear circuits area the situation is somewhat different because of the absence of basic circuits [Ref. (8), p. 99]. However, this situation is rapidly changing with the development of pertinent linear time-invariant circuit synthesis methods[9–12] using presently available integrated capacitors [Ref. (13), pp. 247–255], integrated operational amplifiers,[14] and integrated variable gyrators.[15, 16] Synthesis methods in terms of negative impedance converters have been described,[17] in conjunction with distributed circuit concepts,[17, 18] while very practical active RC lumped-distributed circuitry has been developed.[19] Likewise, the presence of electronically adjustable gyrators has led to apparently practical means of obtaining integrated variable delay lines,[20] voltage controlled tunable structures,[21] and electronically variable oscillators and mixers.[22] Still, although the position of specific integrated electronically tunable structures seems clear, practical synthesis of broad classes of time-variable circuits for integration have not generally been discussed in the literature.

Here we point out the further versatility of integrated circuits by theoretically developing a readily and practically integrated time-variable synthesis of quasilossless networks. The theory rests upon the use of gyrators to obtain all coupling and time-variable constraints; its practical significance lies in the recent integration of the gyrator[23] through a commercially reproducible structure.

The theory rests upon the Spaulding decomposition [Ref. (24), p. 6, (25)] of finite lossless time-domain immittances. In actual fact this decomposition holds for the more general class of quasilossless immittances, defined by Anderson [Ref. (26), p. 41] as an immittance of any circuit connexion of a finite number of passive lossless circuit elements (i.e. variable transformers and gyrators and fixed inductors and capacitors). The quasilossless synthesis to be described uses only three-terminal, grounded, time-variable gyrators and a minimum number of fixed capacitors, also grounded. Although a synthesis could be obtained by making cascade gyrator replacements(27) of all transformers in previously obtained structures,(28, 25, 29) the results would generally be impractical because of the shorting of elements by the common ground required in presently obtainable integrated gyrators; an overly excess number of gyrators would also be needed. Nevertheless, by introduction of extra circuitry the gyrators can be ungrounded if so desired,(30–32) though this apparently less practical approach will not be taken here. Consequently, a different interpretation is developed here for the Spaulding decomposition of an $n \times n$ time domain impulsive-admittance $\mathbf{y}(t,\tau)$. This interpretation obtains a ground common to all gyrators and capacitors allowing for proper biasing of the transistors (and diodes) realizing the gyrators (and capacitors) without disruption of circuit operation.

2. PRELIMINARIES

In this section we quickly review for ready reference the background notions needed for synthesis, most of which are rigorously treated in the somewhat unavailable references (24), (26), (33), (34).

To be perfectly rigorous and to avoid paradoxical pitfalls it is suitable to consider an n-port network's port voltages $\mathbf{v}$ and currents $\mathbf{i}$ as n-vectors of infinitely differentiable functions of time, t, which are zero until some finite time. If we let a superscript tilde denote matrix transposition, for a given network $\mathbf{N}$ the input energy

$$\mathcal{E}(t) = \int_{-\infty}^{t} \tilde{\mathbf{v}}(\tau)\mathbf{i}(\tau)\mathrm{d}\tau \qquad (2.1)$$

is then well-defined for all pairs $[\mathbf{v}, \mathbf{i}]$ of voltage and current allowed by the network at its ports; we write $[\mathbf{v}, \mathbf{i}] \in \mathbf{N}$ for such allowed pairs [Ref. (35), p. 8]. If $\mathcal{E}(t) \geqslant 0$ for all $[\mathbf{v}, \mathbf{i}] \in \mathbf{N}$ and all finite t, then $\mathbf{N}$ is called *passive*. If arbitrary square-integrable voltage sources $\mathbf{e}$ of one ohm internal source impedances are connected to a passive $\mathbf{N}$ and if $\mathcal{E}(\infty) = 0$ for all resulting $[\mathbf{v}, \mathbf{i}] \in \mathbf{N}$, then $\mathbf{N}$ is called *lossless*. A passive network constructed from a finite number of time-variable transformers, time-variable gyrators, and time-invariant, passive, inductors and capacitors is called *quasilossless*. Although the circuit elements in a linear quasilossless network are individually lossless their interconnexion need not be lossless;(36) this is seen, for example, by a unit capacitor loaded transformer whose time-variable turns ratio drops at some finite time to, and remains thereafter at, zero. Still, lossless networks are special cases of quasilossless ones for which the necessary and sufficient synthesis conditions for realizability in the linear case are known in terms of the various network descriptions.

We will assume that the network $\mathbf{N}$ is described by its *impulsive admittance* $\mathbf{y}(t,\tau)$, this being defined through

$$\mathbf{i}(t) = \int_{\infty}^{\infty} \mathbf{y}(t,\tau)\mathbf{v}(\tau)\mathrm{d}\tau \qquad (2.2)$$

for all $[\mathbf{v}, \mathbf{i}] \in \mathbf{N}$. However, the integral of (2.2) must be rigorously interpreted in the distribution theory sense with $\mathbf{y}$ a distributional kernel, the "physical" impulse response matrix. For conciseness (2.2) is written as $\mathbf{i} = \mathbf{y} \cdot \mathbf{v}$; the impulsive impedance $\mathbf{z}(t,\tau)$ is, therefore, defined by $\mathbf{v} = \mathbf{z} \cdot \mathbf{i}$. When, as is true in the quasilossless case, $\mathbf{N}$ is *finite*, that is, described by differential operators $\mathbf{a}(.\,,.)$ and $\mathbf{b}(.\,,.)$ through

$$\mathbf{a}(p,t)\mathbf{v}(t) = \mathbf{b}(p,t)\mathbf{i}(t), \quad p = \mathrm{d}/\mathrm{d}t \qquad (2.3)$$

then the entries of $\mathbf{y}$ are antecedal (i.e. nonanticipatory) solutions of the describing equations (2.3), subject to impulsive input voltages. Corresponding to (2.3) is an adjoint set of equations having the anticipative (i.e. nonantecedal) impulse response matrix $\mathbf{y}^a$ defined by $\mathbf{y}^a(t,\tau) = \tilde{\mathbf{y}}(\tau,t)$. As with all sets of differential equations, this adjoint set has an equally valid antecedal impulse response matrix $\mathbf{y}^a_a$. The antecedal adjoint $\mathbf{y}^a_a$ is of considerable interest since, when subject also to the passivity condition of (2.1), the *quasilossless condition* is [Refs. (25), (26), p. 41]

$$\mathbf{y} + \mathbf{y}^a_a = \mathbf{0}_n \qquad (2.4)$$

where $\mathbf{0}_n$ is the $n \times n$ zero matrix. An admittance (or impedance) having $\mathcal{E}(t) \geqslant 0$ and satisfying (2.4) will, as a consequence, be called *quasilossless*.

A system of differential equations (2.3) has an antecedal impulsive admittance [Refs. (24), p. 86; (37), p. 184; (38), p. 529]

$$\mathbf{y}(t,\tau) = \sum_{j=0}^{r} \mathbf{y}_j(t)\delta^{(j)}(t-\tau) + \tilde{\mathbf{\Psi}}(t)\Theta(\tau)u(t-\tau) \qquad (2.5)$$

where u is the unit step function, δ is the unit impulse with $\delta^{(j)}$ its jth derivative, the $\mathbf{y}_j$ are $n \times n$ matrices while $\mathbf{\Psi}$ and Θ are $q \times n$ matrices whose q rows are linearly independent over all time [Ref. (38), p. 530]. The passivity requirement shows, by somewhat tricky reasoning [Ref. (24), pp. 47–51], that $\mathbf{y}_j = \mathbf{0}_n$ for $j \geqslant 2$ and that $\mathbf{y}_1$ is symmetric and positive semidefinite [Ref. (39), p. 8–4; (24), p. 52]. Application of (2.4) then shows that one can obtain the following *Spaulding decomposition* of a quasilossless impulsive admittance [Ref. (24), p. 67]

$$\mathbf{y}(t,\tau) = \tilde{\mathbf{N}}(t)\{\delta^{(1)}(t-\tau)\mathbf{1}_l\}\mathbf{N}(\tau) + \mathbf{y}_0(t)\delta(t-\tau) + \tilde{\mathbf{M}}(t)\{u(t-\tau)\mathbf{1}_c\}\mathbf{M}(\tau) \qquad (2.6)$$

where $\mathbf{1}_c$ and $\mathbf{1}_l$ are the identity matrices of order c and l with $l = [\text{rank } \mathbf{y}_1] \leqslant n$ and $\mathbf{y}_0 = -\tilde{\mathbf{y}}_0$; of course $c \leqslant q = [\text{number of rows of } \Theta]$, and c will be chosen the smallest possible. On physical grounds,[40] and for mathematical simplicity, we will assume the entries of $\mathbf{N}$, $\mathbf{M}$, and $\mathbf{y}_0$ to be infinitely differentiable functions.

3. QUASILOSSLESS SYNTHESIS

We begin by noting that it is a somewhat straightforward matter[15, 23] to construct an integrated, time-variable, 3-terminal, grounded gyrator of gyration conductance $\gamma(t)$. Such is symbolized by Fig. 1 and described by

$$\mathbf{y}(t,\tau) = \begin{bmatrix} 0 & -\gamma(t) \\ \gamma(t) & 0 \end{bmatrix} \delta(t-\tau) \qquad (3.1)$$

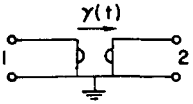

FIG. 1. Grounded gyrator.

Any $n \times n$ matrix of the form

$$\mathbf{y}(t,\tau) = \mathbf{y}_0(t)\delta(t-\tau), \qquad \mathbf{y}_0 = -\tilde{\mathbf{y}}_0 \qquad (3.2)$$

is then easily realized by at most $n(n-1)/2$ gyrators. For this realization one merely writes $\mathbf{y}_0$ as the sum of skew-symmetric matrices of rank two, each such matrix corresponding to one gyrator, interconnecting two of the ports, and open circuits, all ports having one common grounded terminal. The subnetworks realizing the rank two matrices are then connected in parallel to obtain $\mathbf{y}_0(t)\delta(t-\tau)$. For example, for

$$\mathbf{y}_0 = \begin{bmatrix} 0 & -\gamma_1 & -\gamma_2 \\ \gamma_1 & 0 & -\gamma_3 \\ \gamma_2 & \gamma_3 & 0 \end{bmatrix} = \begin{bmatrix} 0 & -\gamma_1 & 0 \\ \gamma_1 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} + \begin{bmatrix} 0 & 0 & -\gamma_2 \\ 0 & 0 & 0 \\ \gamma_2 & 0 & 0 \end{bmatrix} + \begin{bmatrix} 0 & 0 & 0 \\ 0 & 0 & -\gamma_3 \\ 0 & \gamma_3 & 0 \end{bmatrix} \qquad (3.3)$$

the realization of $\mathbf{y}_0(t)\delta(t-\tau)$ is as shown in Fig. 2. As a consequence, the $(n+m) \times (n+m)$ matrix

$$\mathbf{y}_c(t,\tau) = \begin{bmatrix} \mathbf{0}_n & -\tilde{\mathbf{\Gamma}}(t) \\ \mathbf{\Gamma}(t) & \mathbf{0}_m \end{bmatrix} \delta(t-\tau) \qquad (3.4$$

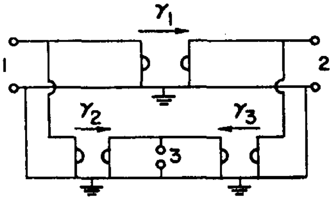

FIG. 2. 3-port $\mathbf{y}_0\delta$ realization.

is also readily realized by at most nm grounded gyrators. For notational convenience the described gyrator connexion realizing (3.2) will be drawn as in Fig. 3(a) while that for $\mathbf{y}_c$ of (3.4) is shown in Fig. 3(b).

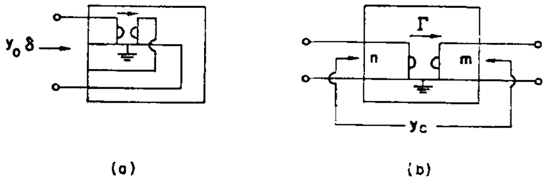

FIG. 3. Gyrator realization conventions.

If now we consider Fig. 4 where an m-port described by its impulsive impedance $\mathbf{z}_l$ cascade loads a gyrator $(n+m)$-port described by (3.4), then we find, from $\mathbf{v}_2 = -\mathbf{z}_l \cdot \mathbf{i}_2$ and $\begin{bmatrix} \mathbf{i} \\ \mathbf{i}_2 \end{bmatrix} = \mathbf{y}_c \cdot \begin{bmatrix} \mathbf{v} \\ \mathbf{v}_2 \end{bmatrix}$

$$\mathbf{y}(t,\tau) = \tilde{\Gamma}(t)\mathbf{z}_l(t,\tau)\Gamma(\tau) \tag{3.5}$$

Note that, as expected, the gyrators change a load into its dual.

grounded gyrators of Fig. 3(b), with this connexion in fact placing the same ground upon the capacitors as on the gyrators. The $\tilde{\mathbf{N}}\{\delta^{(1)}\mathbf{1}_l\}\mathbf{N}$ term represents l unit inductors loading grounded gyrators; the inductors can in turn be realized by loading a gyrator $2l$-port having $\Gamma = \mathbf{1}_l$ by l unit capacitors, as the relation dual to (3.5) shows. These networks realizing the three terms of (2.6) are then connected in parallel to obtain the final structure shown in Fig. 5. It should be observed that the same ground is carried common throughout the structure.

One sees, somewhat intuitively, that the process uses the minimum number of capacitors since the state space dimension for any realization of (2.6) must be at least $c+l$, and we have achieved dimension $c+l$.

As an example to illustrate the procedure consider

$$\mathbf{y}(t,\tau) = \begin{bmatrix} 0 & -t\sin t \\ t\sin t & 0 \end{bmatrix}\delta(t-\tau)$$
$$+ \begin{bmatrix} \cos 2t\cos 2\tau + \sin t\sin\tau & 2\cos 2t\cos 2\tau + 3\sin t\sin\tau \\ 2\cos 2t\cos 2\tau + 3\sin t\sin\tau & 4\cos 2t\cos 2\tau + 9\sin t\sin\tau \end{bmatrix} u(t-\tau) \tag{3.6a}$$

for which the Spaulding decomposition is

$$\mathbf{y}(t,\tau) = \begin{bmatrix} 0 & -t\sin t \\ t\sin t & 0 \end{bmatrix}\delta(t-\tau) + \begin{bmatrix} \cos 2t & \sin t & 0 \\ 2\cos 2t & 3\sin t & \cos 2t \end{bmatrix}$$
$$\times \begin{bmatrix} u(t-\tau) & 0 & 0 \\ 0 & u(t-\tau) & 0 \\ 0 & 0 & u(t-\tau) \end{bmatrix} \begin{bmatrix} \cos 2\tau & 2\cos 2\tau \\ \sin\tau & 3\sin\tau \\ 0 & \cos 2\tau \end{bmatrix} \tag{3.6b}$$

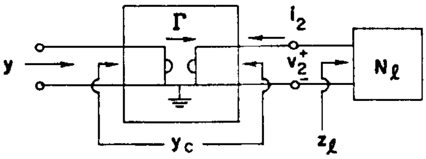

FIG. 4. Cascade load connexion.

At this point we can realize a given quasilossless impulsive admittance $\mathbf{y}(t,\tau)$ by directly interpreting Spaulding's decomposition (2.6). The realization of $\mathbf{y}_0(t)\delta(t-\tau)$ has already been described. The term $\mathbf{M}\{u\mathbf{1}_c\}\mathbf{M}$ represents c unit capacitors loading the

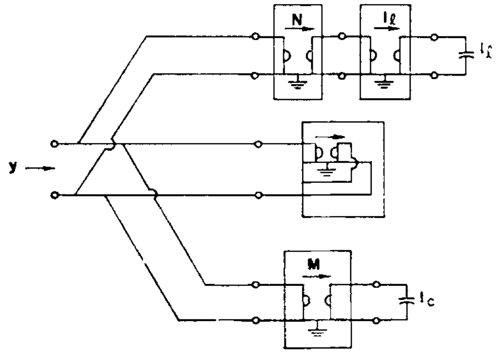

FIG. 5. Grounded capacitor gyrator admittance realization.

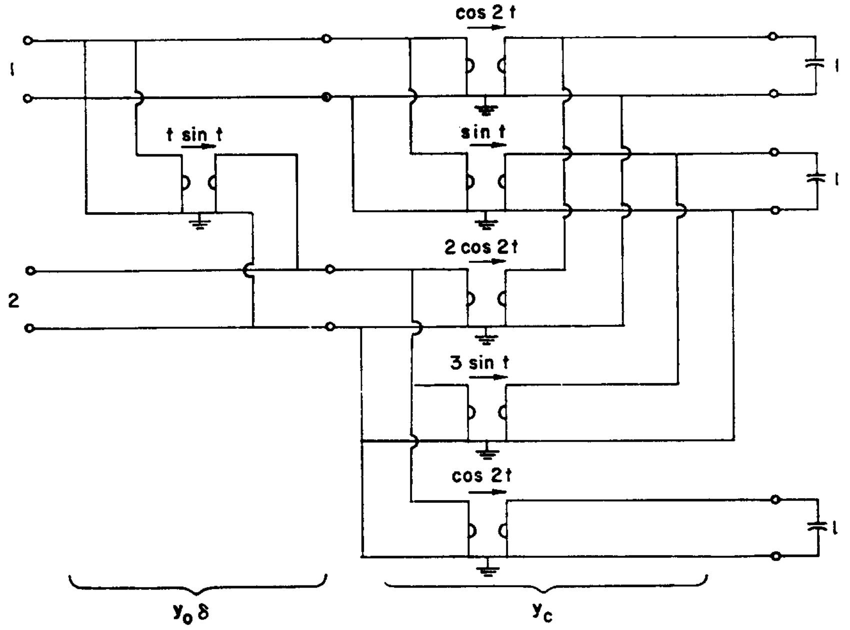

FIG. 6. Example 2-port realization.

From (3.6b) the quasilossless network of Fig. 6 results.

One practical problem shows up in that the gyrators of Fig. 6 specifically, and Fig. 5 generally, have gyration conductances which change signs. This problem can be solved[41] by connecting two gyrators in parallel, one to realize the positive and one the negative portion of the gyration conductance; a negative gyration conductance is obtained from a positive one by reversing the ports (when $\gamma > 0$ the gyrator in the parallel combination for the negative portion is adjusted to be an open circuit, for example).

4. TIME-INVARIANT SYNTHESIS

In the time-invariant case, where $\mathbf{y}(t,\tau) = \mathbf{y}(t-\tau,0)$, the procedure of the last sections yields a synthesis generally incorporating non time-invariant elements. To eliminate these one can make judicious combinations of variable elements,[42] however, it is simpler to proceed directly to obtain a time-invariant grounded gyrator minimal (grounded) capacitor structure. For this we consider the Laplace transform characterization, $\mathbf{Y}(p) = \mathscr{L}[\mathbf{y}(t,0)]$, which must be positive-real. Since, in the time-invariant case, a quasilossless network must be lossless (as a finite connexion of fixed lossless elements [Ref. (35), p. 153]), then [Ref. (35), p. 102] $\mathbf{Y}(p) = -\tilde{\mathbf{Y}}(-p)$ and the standard lossless partial-fraction decomposition holds [Ref. (35), p. 128]

$$\mathbf{Y}(p) = \mathbf{B}_\infty + p\mathbf{A}_\infty + p^{-1}\mathbf{A}_0 + \sum_{i=1}^{r} \mathbf{Y}_i(p); \quad \mathbf{Y}_i(p) = \frac{p\mathbf{A}_i + \mathbf{B}_i}{p^2 + \omega_i^2} \tag{4.1}$$

where each of the terms in the sum is individually positive-real and lossless. We can then write [Ref. (35), pp. 205, 210]

$$p\mathbf{A}_\infty = \tilde{\mathbf{T}}_l[p\mathbf{1}_l]\mathbf{T}_l \tag{4.2a}$$

$$p^{-1}\mathbf{A}_0 = \tilde{\mathbf{T}}_c[p^{-1}\mathbf{1}_c]\mathbf{T}_c \tag{4.2b}$$

$$\frac{p\mathbf{A}_i+\mathbf{B}_i}{p^2+\omega_i^2}=\sum_{j_i=1}^{n}\tilde{\mathbf{T}}_{j_i}\mathbf{Z}_i(p)\mathbf{T}_{j_i};\quad \mathbf{Z}_i^{-1}(p)=\begin{bmatrix}p & -\omega_i\\ \omega_i & p\end{bmatrix} \tag{4.2c}$$

where $\mathbf{T}_l$, $\mathbf{T}_c$, and $\mathbf{T}_{j_i}$ are constant, generally non-square, matrices; the first j_i-1 columns of the $2\times n$ matrices $\mathbf{T}_{j_i}$ can in fact be chosen zero [Ref. (35), p. 210]. The procedure for realization of $\mathbf{Y}(p)$ is the same as for $\mathbf{y}(t,\tau)$ of Fig. 5, since the p-plane equivalent of (3.5) holds as applied to (4.2), except that $\mathbf{Z}_i(p)$ is obtained as the 2-port shown in Fig. 7. Since the procedure uses [degree of $\mathbf{Y}(p)$] capacitors, it uses the minimum number of capacitors.

As an example, consider the Brune section of Fig. 8, which is of use in cascade synthesis[9] and which has

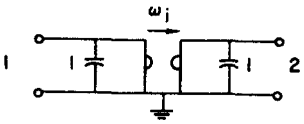

FIG. 7. $\mathbf{Z}_i(p)$ realization.

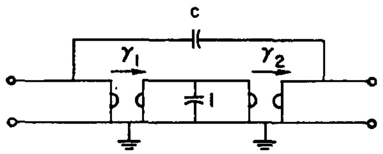

FIG. 8. Non-reciprocal Brune section.

$$\mathbf{Y}(p)=cp\begin{bmatrix}1 & -1\\ -1 & 1\end{bmatrix}+\frac{1}{p}\begin{bmatrix}\gamma_1^2 & -\gamma_1\gamma_2\\ -\gamma_1\gamma_2 & \gamma_2^2\end{bmatrix} \tag{4.3a}$$

A grounded capacitor equivalent arises from writing

$$\mathbf{Y}(p)=\begin{bmatrix}1\\ -1\end{bmatrix}[cp]\,[1\ \ -1]+ + \begin{bmatrix}\gamma_1\\ -\gamma_2\end{bmatrix}\left[\frac{1}{p}\right][\gamma_1\ \ -\gamma_2] \tag{4.3b}$$

from which two respective coupling admittances are derived, as for (3.4) and (3.5),

$$\mathbf{Y}_c=\left[\begin{array}{cc|c}0 & 0 & -1\\ 0 & 0 & 1\\ \hline 1 & -1 & 0\end{array}\right]\quad \mathbf{Y}_{c_s}=\left[\begin{array}{cc|c}0 & 0 & -\gamma_1\\ 0 & 0 & \gamma_2\\ \hline \gamma_1 & -\gamma_2 & 0\end{array}\right] \tag{4.3c}$$

The grounded capacitor (and gyrator) equivalent is shown in Fig. 9. One can proceed similarly for the Darlington sections.[9]

5. DISCUSSION

Using a method of grounded gyrator imbedding of capacitors a synthesis of (passive) quasilossless impulsive admittances has been given. The method uses a minimum number of capacitors, all of which

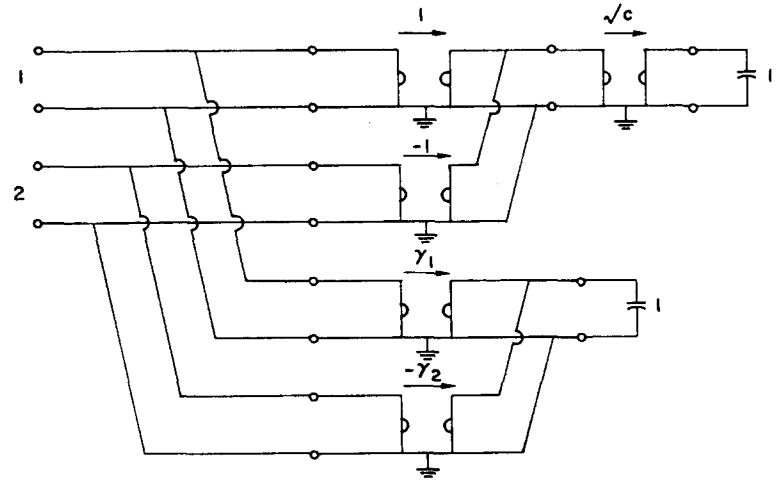

FIG. 9. Grounded capacitor gyrator Brune section.

are grounded, though a more rigorous proof of this minimality would be desirable for the time-variable case. All capacitors, which are time-invariant, have been assumed equal for the convenience of integrated circuit designs; however, any non-zero values can be obtained without any structural change by straightforward denormalization.

As the equivalent circuits of Figs. 8 and 9 show, more than the minimum number of gyrators may be used even though this number is often fewer than needed for simple transformer replacements. But idealized gyrators appear to be more suitably obtained in integrated form than are idealized capacitors so the synthesis is quite practically oriented in this regard. Consequently, when equivalences of the type of Figs. 8 and 9 are used, a previous cascade filter synthesis[9] should become even more practical for integrated circuit design even though several more gyrators are introduced. Because of this filter synthesis the time-invariant portion of the paper may prove the most immediately useful especially when adjustable filters are considered.

It should be observed that even though n-port methods have been used the resulting structure is an $(n-1)$-terminal network; the theory is well justified by the Cauer equivalence between n-ports and $(n-1)$-terminal networks [Ref. (35), p. 17].

Because of the physical layout of thin film devices, as compared to monolithic ones, one feels that appropriate development of the thin film technology[43, 44] can take most advantage of the methods presented. This especially since the thin film transistors[45] are particularly suited for the variations needed[16] in obtaining variable gyrators. However, the material is directly applicable to monolithic structures where the various separate components have been constructed at Stanford.

Given a quasilossless impedance $\mathbf{z}(t,\tau)$ or scattering matrix $\mathbf{s}(t,\tau)$, a synthesis can proceed by conversion to $\mathbf{y}(t,\tau)$ when the latter exists. However, as examples show,[46] the three matrices $\mathbf{s}$, $\mathbf{y}$, and $\mathbf{z}$ can independently exist.

For nonquasilossless networks, synthesis methods are presently under development which allow any realizable $\mathbf{y}$ to result from loading a quasilossless structure by resistors. Consequently, the results of this paper appear basic to any rigorous theory of integrated circuits.

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