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Active and Periodically-Time-Varying Filters in Sonar

W. Saraga
The General Electric Company Limited, Telecommunications Research Laboratories, Hirst Research Centre, Wembley, England

1. Introduction This paper is concerned with linear electric networks used for the processing of analogue signals in sonar equipment. The processing requirements are in general similar to those occurring in telecommunication systems; i.e. filters, delay, signal shaping and equalising networks, with highly frequency-selective transfer functions T(p) (p being the complex frequency parameter), are required. There is, however, in the case of sonar a restriction on space which can be met only by microelectronic realisation of the signal processing networks.

Conventionally, resistor-inductor-capacitor (RLC) circuits would be used. However, inductors cannot be realised in microelectronic form and RC networks on their own cannot produce highly selective characteristics, like filter responses with sharp cut-offs or all-pass responses with non-monotonic delay-frequency curves. In order to achieve both high selectivity and microelectronic realisability we have, in addition to R and C, to use transistors (T) as circuit elements (or similar semiconductor devices).

2. Network Theoretical Considerations The high selectivity which can be obtained with RLC filters, is achieved by interaction of positive reactances (not available in the case of passive and time-invariant RC networks) and negative reactances and by placing some of the poles of T(p) very near the imaginary p-axis (in the RC case the poles are restricted to the negative real p-axis). In RCT networks these restrictions can be overcome by various methods, none of which are fully satisfactory, because the sensitivity of the performance characteristics of RCT networks is in many (although not in all) respects significantly higher than that of corresponding RLC networks. Therefore, research aimed at sensitivity minimisation, supplemented by component research to minimise their variability, is actively pursued at many places.

Three main approaches to the synthesis of highly selective RCT networks are being used: by non-reciprocity, by activity, by the use of periodically time-varying parameters. It is relevant, in practice, that non-reciprocity - although in theory compatible with passivity, can in practice only be realised with the help of activity (periodically time-varying networks can often be simplified by inclusion of active components).

3. Non-Reciprocal Methods The simplest non-reciprocal element is the gyrator (G) which is a passive two-port (its realisation is discussed below) with transmission matrix $\begin{bmatrix} 0 & R \\ R^{-1} & 0 \end{bmatrix}$ so that $v_1=-Ri_2$, $v_2=Ri_1$ (see Fig. 1a). If port 2 is terminated by an impedance Z_2, then

the input impedance at port 1 is $Z_1=R^2/Z_2$. Thus, if port 2 is terminated by a capacitance C, we obtain $Z_1 = pR^2C$, i.e. the impedance of an inductance $L=R^2C$. Therefore, with the help of gyrators, we can simulate RLC networks (and utilise existing RLC designs) by simulating each L by a GC combination. If none of the terminals of a required L is earthed, a "floating" gyrator is required, which is feasible but inconvenient. In this case the equivalence shown in Fig. 1b can be utilised (easily verifiable by matrix multiplication)(1).

Gyrators can be simulated electronically in various ways. Fig. 2a shows a suitable circuit consisting of two ideal voltage controlled current sources; Fig. 2b shows a gyrator circuit (input port (1,0); output port (2,0)) containing three "ideal" transistors (the collector current being assumed to be equal to the emitter current and the emitter voltage equal to the base voltage) and two impedances $Z_a=Z_b=R$ (proof by inspection).

4. Active Methods There are various methods for utilising activity to overcome the limitations of passive RC networks, only some of which will be discussed.

4.1. Simulating methods These are similar to those discussed in Section 3. If the gyrator in Fig. 2b is terminated at port (2,0) by a capacitor, and if port (3,0) is regarded as the output port (Z_b being removed), then we obtain a generalised impedance converter (GIC), which in this particular case converts a resistance into an inductance. Floating inductors can also be simulated by means of two such convertors (see Fig. 2c)(2). This method can be generalised to simulate the L-subnetwork of an RLC network by an R-subnetwork, connected via a multi-GIC network to the RC-subnetwork.

A related method(3) is based on the fact that we can obtain a one-port with impedance $Z=1/(RC^2p^2)$, for example by choosing in Fig. 2b $Z_a=Z_b=1/(pC)$ and terminating port (2,0) in R. This permits us to replace in an RLC network each element with impedance Z by an element with impedance Z/p, which leaves the voltage transfer ratio unchanged; the transformed network can then be realised as an active RC circuit.

4.2. Feedback methods Only principles will be discussed. In the circuit in Fig. 3a, where $T(p)=N(p)/D(p)$ and $N(p)$, $D(p)$, $D_o(p)$ are polynomials in p, the denominator $D_o(p)$ of the overall transfer function $T_o(p)$, $D_o=D-A\beta N$, is not subject to the restrictions applying to $D(p)$. Typical practical circuits(4a,b;5) of this kind, where $T(p)$ and β are interdependent, are shown in Figs. 3b,c. A different principle is involved in the circuit in Fig. 4a for $\mu \to \infty$; its overall transfer function approaches $T_o(p)=1/T(p)$. Since only the poles, but not the zeros of $T(p)$ are restricted to the negative real p-axis, the poles of $T_o(p)$ are derestricted. By cascading this circuit with an RC circuit with transfer function $N_1(p)/D(p)$, we can obtain an arbitrary transfer function $N_1(p)/N(p)$. An example is shown in Fig. 4b.

Whereas by the methods described in Section 4.1. the required T(p) is realised as a whole, the feedback methods are usually applied to quadratic and/or biquadratic functions, into which T(p) is factorised, the corresponding circuits being cascaded via buffer amplifiers. Particularly low sensitivity is obtained by a 3-amplifier circuit (see Fig. 5) simulating an RLC resonant circuit(6).

5. Periodically-Time-Varying Methods The basic principle of these methods is to carry out the filtering operation in a much lower frequency band than that in which the selection is required. In this way RC networks can be used for the filtering operation, whereas in the original frequency band RLC or even crystal networks would be required. Two stages of frequency translation by modulation are involved, and at least two paths are required (Fig. 6a) in order to

suppress, by mutual cancellation, output frequency components not contained in the input. A generalisation of this concept is represented by the N-path filter[7] (Fig. 6b). Band-limiting input and output filters ($f<\frac{1}{2}Nf_c$, f_c=carrier frequency) ensure that a transfer function exists, i.e. any input frequency produces only an identical output frequency. Instead of 2N modulators only two commutating switches are needed which can be realised as microelectronic MOST devices. In terms of the transfer function H(p) of the (identical) path networks the overall transfer function T(p) is proportional to $e^{j\gamma}H(p+j\omega_c)+e^{-j\gamma}H(p-j\omega_c)$. Thus, if H(p) is an RC transfer function so that $H\to\infty$ if $p=-\sigma_i$ (σ_i real, >0), then $T\to\infty$, if $p=-\sigma_i\pm j\omega_c$; i.e. the restriction applying to the poles of time-invariant RC networks does not apply to periodically time-varying RC networks.

Single sideband generation and detection by quadrature modulation is based on a somewhat similar principle: highly selective discrimination between the sidebands is obtained by signal processing with RC networks at base band frequencies. A microelectronic realisation of a demodulator of this type (involving also active filters) is described in ref. 8.

6. Sample-and-Hold Circuits (time-varying active RC circuits) Any required function T(p) can be synthesised by tapping a multi-section delay line at its junctions and feeding the sums of appropriately weighted junction voltages to the input and the output of the line[9]. The delay sections can be realised by gated capacitance stores controlled by a clock frequency the delay being proportional to this period (see Fig. 7). Ref. 10 describes a microelectronic realisation of such a delay circuit, using tantalum thin film capacitors and resistors, a monolithic m.o.s.t. circuit having two 3-way switches and an operational amplifier.

7. Sensitivity Pass band sensitivity of feedback circuits increases with the Q-values of the quadratic factors of the denominator of T(p), and is much smaller for the 3-amplifier circuit than for single amplifier circuits. Realisation by simulation usually leads to less sensitivity than realisation by factorisation and buffering[11]. Time-varying circuits usually have much lower pass band sensitivity (often lower than that of RLC circuits) but increased stop band sensitivity, and/or the suppression of frequency components not contained in the input is highly sensitive. It is relevant to note, that our final practical aim is not minimisation of sensitivity as such, but minimisation of total variation of response.

8. Other Requirements Once the sensitivity problem has been solved, other points like component value spread, noise, dynamic range, heat dissipation, and ultimately reliability and cost require consideration.

9. Practical Results and Conclusion Much of the relevant work being done at present is still in the stage of feasibility studies[12]. However, practical microelectronic circuits have been developed which are applicable to narrow band pass filters used in the Doppler type of sonar receiver and to delay networks used in beam forming arrangements and will be described at the Meeting.

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