DISTRIBUTED MICROWAVE ACTIVE FILTERS WITH GaAs FETs

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Abstract — The study focuses on the application of transversal and recursive principles to the design of broadband active filters for microwave frequencies, employing microwave transistors as active elements. Physical implementation of these concepts is illustrated with a transversal-type and a recursive-type experimental filter. To offer a convenient basis for comparison, both examples are chosen to provide 9-to-15-GHz bandpass responses.

INTRODUCTION

At very low frequencies, active filters are quite commonly employed to provide physical realization of higher-order transfer functions that are cost-effective and compact in size. Among other attributes, active filters have the ability not only to compensate for parasitic losses affiliated with passive circuit elements, but also to offer overall amplification. With these same considerations equally relevant to applications in the high-frequency regime, it is thus natural to inquire into possible adaptation of low-frequency operational-amplifier-based techniques for meeting filter needs at microwave frequencies. An application area of particular interest relates to microwave monolithic circuit implementations. The transposition process is unfortunately impeded by the lack of suitable broadband, high-gain devices to perform operational amplifier functions. Although the gain issue is a major concern, it is really time delay intrinsic to microwave active devices that most often represents the principal limiting factor. Consequently, most of the sporadic interest in microwave active filters has concentrated on alternative approaches [1,2,3] in which individual reactances and resonators are replaced with microwave active substitutes that yield higher-Q performance.

The present study focuses on distributed active filter solutions which, in general, have the potential of being less constrained by active device gain and time delay considerations. This permits distributed circuits to more readily cope with broad bandwidth requirements, thereby distinguishing such circuits from the lumped-element-type examples mentioned above. The two basic categories of filters to be explored comprise transversal designs and recursive designs. Filters in both categories achieve their frequency-selective responses through interaction among individual signal components of different amplitudes and frequency-dependent phase delays. With exception of parasitic feedback effects, transversal filters rely entirely on feedforward techniques, as illustrated in the flow graph representation of Fig. 1a. Recursive filters, on the other hand, specifically utilize feedback, often in conjunction with transversal segments. A corresponding flow graph example is given in Fig. 1b. Availability of feedback is a powerful asset which can translate into compact circuit realizations with superior performance.

The task dealt with in the following is to translate transversal and recursive principles into practical microwave circuit configurations, with implied emphasis on broadband applications. Deviations from strict interpretations of these principles will be permitted to enhance design flexibility and achieve optimum overall solutions. One particular option involves the inclusion of conventional filter segments to perform specific functions. Illustrations of this hybrid approach will be provided by two experimental filter examples.

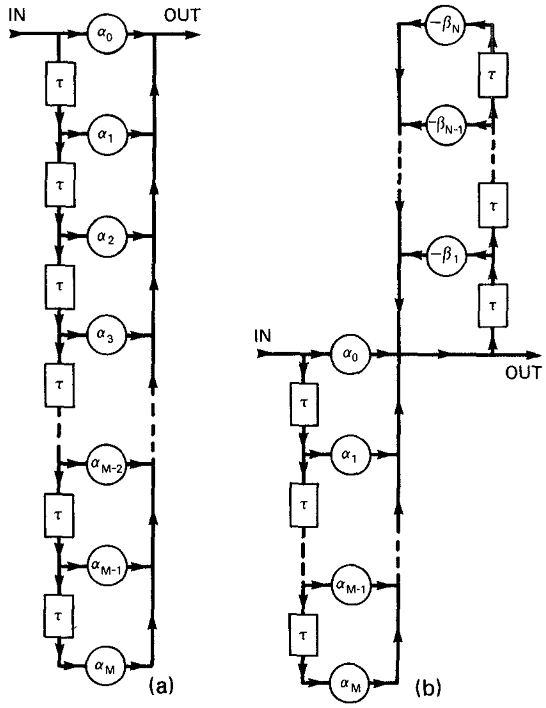

Fig. 1 — Flow graph examples: (a) transversal filter and (b) recursive filter.

THE TRANSVERSAL APPROACH

General Approach — Transversal filters of classical design are closely patterned after their respective flow graph representations (Fig. 1a), comprising delay lines with taps at equal intervals, weighting elements for realizing specific amplitude distributions among the various delayed signal components, and provisions for combining these components to form the output response. The transfer function of a transversal filter with $M + 1$ weighting elements, in terms of the angular frequency ω, is represented by

$$H_T(j\omega) = \sum_{m=0}^{M} \alpha_m \cdot e^{-j2\pi m\omega/\omega_s}, \qquad (1)$$

with coefficients α_m the amplitude weighting factors, and ω_s the symmetry frequency at which the transfer function repeats itself. The delay increment τ introduced in Fig. 1a is related to ω_s through $\tau = 2\pi/\omega_s$. To accommodate the more general situation, in which the basic transversal structure is augmented by conventional filter segments, the composite filter response is written as

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$$H(j\omega) = H_S(j\omega) \cdot H_T(j\omega). \qquad (2)$$

It is thereby assumed that the combined effects of the supplementary filter elements can be condensed into the transfer function $H_S(\omega)$ and subsequently factored out. The decomposition implies that all transversal signal components are equally exposed to these effects—an assumption which is normally valid in most practical situations.

The basic design task is to determine the combination of ω_s and coefficients α_m, $m = 0, 1, \ldots, M$, which permits $H(j\omega)$ to best approximate a prescribed target response function $G(j\omega)$, assumed to represent a bandpass filter for which only the magnitude response matters. Interpreting expression (1) as a truncated Fourier expansion then leads to

$$\alpha_m = \frac{1}{\omega_s} \cdot \int_{\omega_A}^{\omega_B} \frac{|G(j\omega)|}{|H_S(j\omega)|} \cdot e^{+j2\pi(m-M/2)\omega/\omega_s} \cdot d\omega, \qquad (3)$$

with $m = 0, 1, \ldots, M$, and ω_A and ω_B corresponding to the lower and upper bandpass cut-off frequencies, respectively, beyond which $|G(j\omega)|$ vanishes. Both $|H_S(j\omega)|$ and ω_s are considered to be predetermined at this point in the design process. Utilizing discrete Fourier transform yields the alternative expression

$$\alpha_m = \frac{1}{M+1} \cdot \sum_{k=K_A}^{K_B} \frac{|G(j\frac{k}{M+1}\omega_s)|}{|H_S(j\frac{k}{M+1}\omega_s)|} \cdot e^{+j2\pi k(m-M/2)/(M+1)}. \qquad (4)$$

The index limits, K_A and K_B, bracket the passband values of k. For even periodic transfer functions, like the ones to be focused on here, all the coefficients are real, with $\alpha_m = \alpha_{M-m}$ for all m.

Simplifying Approximations — The coefficients invariably exhibit both positive and negative signs. Their realization at microwave frequencies may either involve the use of active devices in inverting and non-inverting configurations, or may rely on a dual feed line approach [4]. The alternative explored here, which primarily pertains to high-pass-type filters with passbands centered around $\omega_s/2$, is to replace all opposite-sign contributions in the sum expression (1), for which $\alpha_k/\alpha_{M/2} < 0$, by approximations according to

$$\alpha_k \cdot e^{-j2\pi k\omega/\omega_s} \approx -\alpha_k \cdot \gamma \cdot \{e^{-j2\pi(k-1)\omega/\omega_s} + e^{-j2\pi(k+1)\omega/\omega_s}\}. \qquad (5)$$

Without loss of generality it is assumed that M is an even number. A value for γ in the vicinity of 0.6 renders above substitutions useful for bandwidths up to an octave. If individual coefficients are small relative to the main transmission terms, it may be permissible to further simplify the design by replacing pairs of symmetrical coefficients with single-sided approximations in which one of the coefficients is deleted and the other assumes a double role. It may be convenient to delete certain non-essential pairs altogether. Systematic exercise of these simplifying approximations leads to a sparse transversal array.

Above substitutions and deletions of course result in some passband deviations from the nominal response, as design parameters yield to constraints. Computer optimization can be employed to adjust the pertinent parameters and bring the filter response back in line with specifications. A repetition of the procedure with a higher value of M might be indicated if there are not enough design variables left to cope with the deviations. At stopband frequencies which lie outside the validity range of approximation (5), the deviations can become more serious. This necessitates the use of a supplementary window function to subdue out-of-band transmission. The solution adopted here is to let the active devices perform the windowing. By selecting appropriate device input and output matching circuitry, active elements can be designed to provide suitable gain roll-off beyond the passband edges, while offering flat amplification within the frequency interval to be controlled by the main transversal process.

Physical Implementation — A number of microwave circuit realizations based on this modified transversal approach were explored in the course of this study. A general observation was that parasitic feedback and residual reflections, when interacting with each other in a circuit comprised of numerous active elements and three-way junctions, can cause undue interference with the nominal transversal process. The solution pursued here is to design isolation properties into the junctions to suppress such interference.

A TRANSVERSAL-TYPE FILTER EXAMPLE

The design objective was to obtain a triple-hump bandpass response between cut-off frequencies 9 and 15 GHz, while minimizing transmission in designated stopband intervals spanning 5 to 8 GHz and 16 to 19 GHz. After considering various alternatives, it was decided to implement the filter as a cascade connection of a passive transversal filter section, a conventional passive filter portion, and a windowing gain section. The transversal portion consists of a three-branch structure in which impedance transforming transmission line segments, asymmetrical single-section Wilkinson splitters and combiners, and a series tap resistor were used to achieve desired amplitude and phase distribution among the three transversal signal components, as well as accomplish isolation between output and input.

The transversal solution is of the distributed high-pass type which provides a passband response centered at $\omega_s/4\pi = 12$ GHz. The filter parameters are derived by calculating a full set of coefficients through application of relationship (4) and than whittling the number of terms down to the three most essential ones with the help of the simplifying approximations outlined previously. As for the passive section, it is included to aid the transversal process in shaping the filter characteristics by introducing zeros of transmission at 6, 8, 16, and 18 GHz. The section consists merely of a parallel connection of two 100Ω transmission lines with lengths of a quarter wave and of five quarter waves at band center, respectively. Finally, the gain section comprises an Avantek M126 GaAs FET in common-source configuration, which is matched at input and output to achieve flat amplification between 8 GHz and 16 GHz, with distinct gain roll-off beyond these points to accommodate the windowing requirements.

The microstrip hardware realization of the composite filter is shown in Fig. 2. The circuit is implemented on a 0.25-mm-thick fiberglass-reinforced Teflon substrate, with 50Ω coaxial external ports. Measured filter response and calculations are compared in Fig. 3, with the transistor biased at half-I_{DSS} and a drain-source voltage $V_{DS} = +3.0$V.

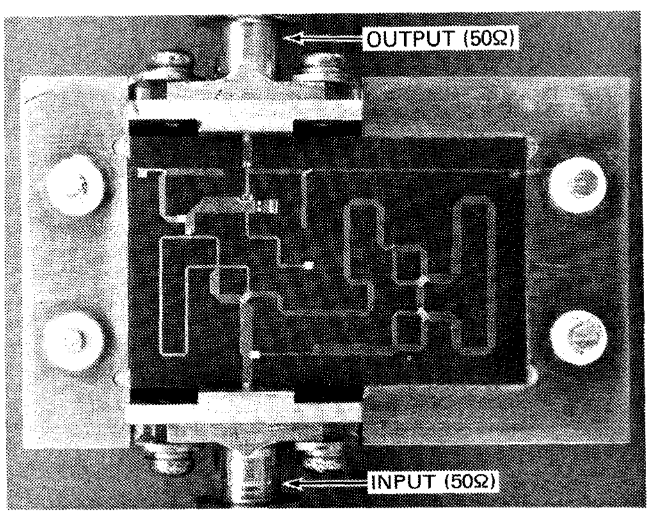

Fig. 2 — Experimental transversal-type bandpass filter.

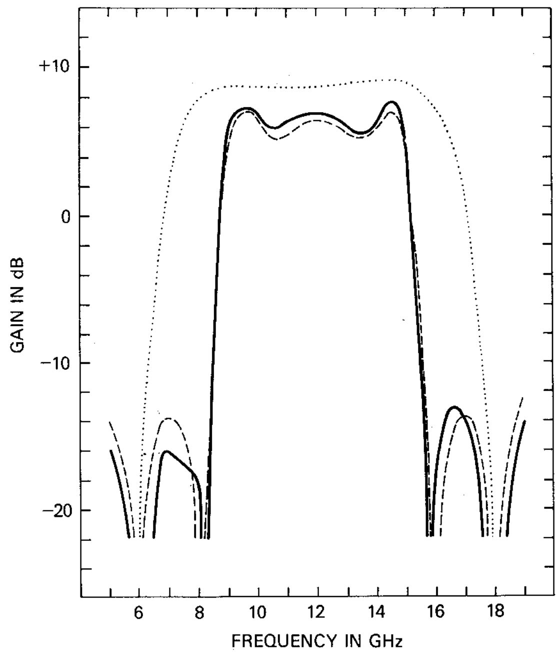

Fig. 3 — Frequency characteristics of the transversal-type experimental filter: ———— measured filter response, ------ calculated filter response, ... calculated characteristics of the amplifier section by itself.

THE RECURSIVE APPROACH

General Approach — Relative to a strictly transversal process, the inclusion of feedback (Fig. 1b) constitutes a very powerful extension, as it permits poles of transmission of arbitrarily high Q-factors to be realized. The transfer function of a recursive circuit with $M+1$ feedforward branches and N feedback branches is given by

$$H_R(j\omega) = \frac{\sum_{m=0}^{M} \alpha_m \cdot e^{-j2\pi m\omega/\omega_s}}{1 + \sum_{n=1}^{N} \beta_n \cdot e^{-j2\pi n\omega/\omega_s}} \tag{6}$$

As in the transversal case, a supplementary transfer function $H_S(j\omega)$ is included to yield a composite filter response $H(j\omega)$ in the factorial form

$$H(j\omega) = H_S(j\omega) \cdot H_R(j\omega). \tag{7}$$

The task is to find the set of coefficient values that yields the best approximation by $H(j\omega)$ to a prescribed target response function $G(j\omega)$. $H_S(j\omega)$ is generally determined by windowing and auxiliary filtering requirements similar to those discussed relative to transversal filters. Of several possible methods for accomplishing this task, two techniques are highlighted in the following.

Impulse Response Method — This approach utilizes conventional s-plane synthesis techniques to approximate the magnitude of the recursive target response $G(j\omega)/H_S(j\omega)$ along the $j\omega$-axis through a rational function approximation $\overline{H}_R(s)$ which may be expressed in terms of partial fractions as

$$\overline{H}_R(s) = \sum_{i=1}^{N_P} \frac{r_i}{s - s_i}, \tag{8}$$

with r_i and s_i representing the residue and the location of the i-th s-plane pole, respectively. An equivalent expression for $\overline{H}_R(s)$ is

$$\overline{H}_R(s) = \sum_{i=1}^{N_P} r_i \int_0^{\infty} e^{(s_i - s)t} \cdot dt \tag{9}$$

from which

$$H_R(s) = \frac{2\pi}{\omega_s} \cdot \sum_{i=1}^{N_P} \frac{r_i}{1 - e^{(s_i - s)2\pi/\omega_s}} \tag{10}$$

can be derived. Through replacement of s by $j\omega$ and rearrangement of exponential terms, expression (10) can be made to conform with definition (6). The α- and β-coefficients then follow by inspection. One of the main drawbacks of this method is that it introduces aliasing effects. This limits application of the approach to functions $\overline{H}_R(j\omega)$ which are sufficiently bandlimited.

Bilinear Transformation Method — The technique adopted here is to obtain the recursive filter solution from a lumped-element low-pass prototype filter. A bilinear transformation is thereby defined which links the frequency variable $j\Omega$ in the lumped-element domain to the frequency variable $j\omega$ in the distributed domain according to some chosen substitution formula $j\Omega = F(j\omega)$. An example of such a formula is represented by

$$F(j\omega) = \Omega_c \cdot \tan \pi\omega_c/\omega_s \cdot \frac{1 + e^{-j2\pi\omega/\omega_s}}{1 - e^{-j2\pi\omega/\omega_s}} \tag{11}$$

which transforms a lumped-element low-pass response with cut-off frequency Ω_c into a distributed high-pass response that exhibits a $2\omega_c$-wide passband centered at $\omega_s/2$. The bilinear transformation results in nonlinear distortion of the prototype frequency axis. In return, no aliasing effects occur. The procedure is to translate the target response for $|H_R(j\omega)|$ given by $|G(j\omega)|/|H_S(j\omega)|$ into the $j\Omega$-domain with the help of the inverse substitution $j\omega = F^{-1}(j\Omega)$. Standard synthesis techniques are then used to approximate the response with a suitable rational expression $\tilde{H}_R(j\Omega)$. Transformation of this expression back into the $j\omega$-domain then yields the solution for $H_R(j\omega)$ in the form of a rational function in exponential terms

$$H_R(j\omega) = \tilde{H}_R(F(j\omega)) = \tilde{H}_R(e^{-j2\pi\omega/\omega_s}) \tag{12}$$

which, again, conforms with format (6).

Physical Implementation — Once the α- and β-coefficients have been determined, approximations analogous to the ones considered earlier for transversal filters may be applied. Of particular interest in the present context are recursive filters which fit the block diagram format shown in Fig. 4. The respective circuits are composed of a chain of generalized two-port transmission elements which are augmented by feedback branches that consist of series-connected impedance elements. The transmission elements TI_n and TO_n, together with optional elements T_{IN} and T_{OUT}, are jointly tasked with performing a variety of functions. One of these functions is, of course to produce the required time delays. A second function concerns the accommodation of supplementary filtering and windowing requirements as described by $H_S(j\omega)$ in expression (7). A requirement of particular relevance to the present case is the realization of transmission zeros, as the degenerate recursive structure in Fig. 4 does not explicitly include feedforward branches to supply these zeros. This approach is based on the observation that reliance on transversal principles to achieve transmission zeros most often constitutes an inefficient use of resources. Finally, in order to impose a sense of direction on the feedback loops, and thereby allow the feedback scheme to operate as intended, the levels of individual signals fed back through the impedance elements must exceed, by a comfortable margin, the levels of parasitic contributions fed forward through the same elements. This is accomplished through assignment of inverting or non-inverting gain functions to selected transmission elements to introduce a sufficient differential between output- and input-related signal levels.

A RECURSIVE-TYPE FILTER EXAMPLE

For the ease of comparison, the same overall objectives were adopted as in the case of the transversal filter example. The design is derived with the help of bilinear transformation (11) which provides passband characteristics centered at $\omega_s/4\pi = 12$ GHz. Special

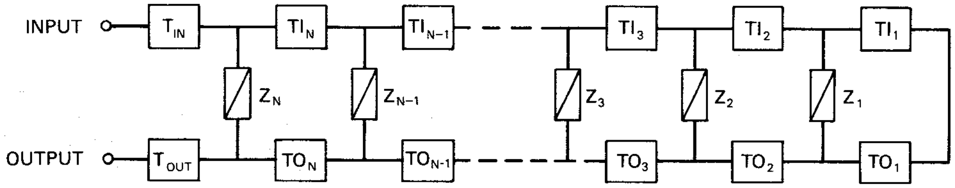

Fig. 4 — Block diagram of a microwave recursive-type active filter.

effort is again focused on achieving the simplest possible solution consistent with the design objectives. Besides adopting the degenerate recursive structure in Fig. 4, all the simplifying approximations and deletions mentioned earlier are utilized. The recursive solution that results, encompassing two feedback loops, also requires involvement of a windowing amplifier section and a passive filter segment to assist in defining the bandpass characteristics. To facilitate comparative evaluation, the gain section and the passive filter, which are both integral parts of the lowest-order feedback loop, are virtually identical to the ones used in the previous example. One of the strengths of this approach is that the intrinsic time delays associated with supplementary components assume constructive roles in the recursive process, laying the foundation for compact filter realizations.

The physical realization of the recursive filter is shown in Fig. 5. The circuit is implemented on a 0.25-mm-thick fiberglass-reinforced Teflon substrate, with 50Ω coaxial input and output ports. The circuit, once more, gets by with only one transistor, an Avantek M126 GaAs FET biased at half-I_{DSS} and at V_{DS} = +3.0V. Figure 6 compares measured and calculated responses.

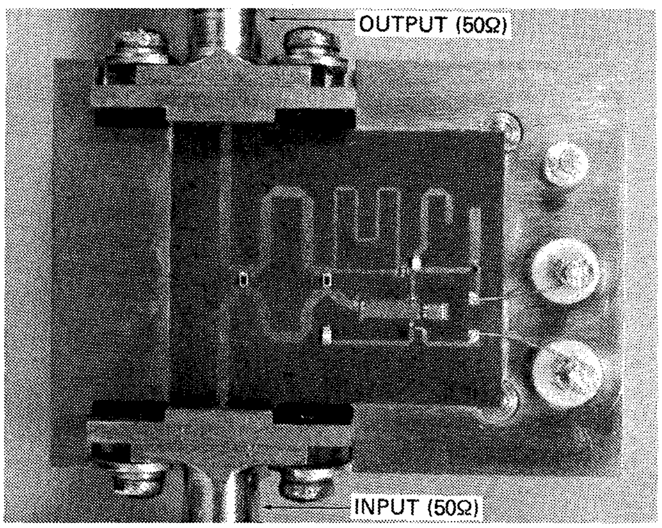

Fig. 5 — Experimental recursive-type bandpass filter.

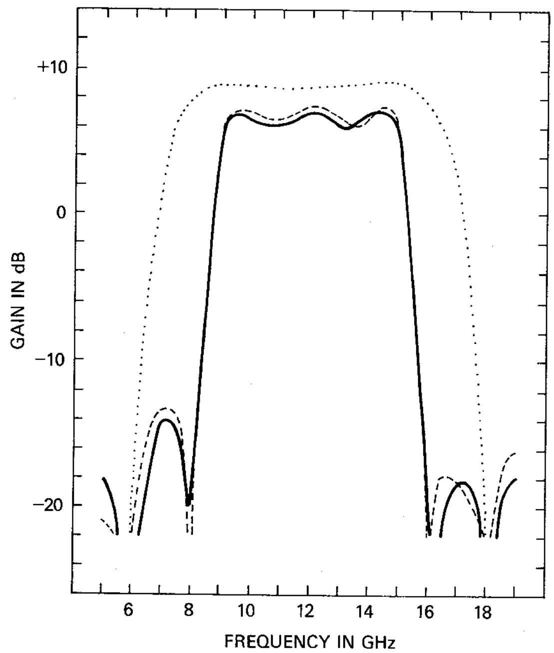

Fig. 6 — Frequency characteristics of the recursive-type active filter example: ——— measured filter response, ------- calculated filter response, ... calculated characteristics of the amplifier section by itself.

CONCLUSIONS

In this study, possibilities for effective use of microwave transistors in broadband filter applications have been explored, with special focus on transversal and recursive concepts and on ways to translate these concepts into practical microwave circuit designs. Based on a comparison between the two approaches, recursive solutions appear to be the favored ones. This is partly due to recursive circuits enjoying the appreciable advantage of available feedback options. They also tend to be more receptive to integration of active devices and supplementary passive filter sections into the overall filter design. Associated time delays can thus be utilized constructively, leading to microwave circuit implementations which are conceptually simple and compact.

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