Comments on "High Selectivity Grounded Capacitor Bandpass and Low-Pass Filters Using the Amplifier Pole"

P. V. ANANDAMOHAN

Abstract—**Three recently proposed grounded capacitor active filters using the amplifier pole have been compared. Some limitations of these circuits for realizing bandpass filters at low frequencies, due to the finite ω_a of the operational amplifier, are discussed. These circuits are also compared with other input-capacitor based bandpass filters.**

In the above letter[1], Nandi described a bandpass/low-pass filter using the operational amplifier pole and grounded capacitor. Nandi's circuit is shown in Fig. 1(a). It has been shown by Nandi that α_1 can be arbitrarily chosen with a given value of C, and the circuit can then be designed to yield the desired pole frequency, pole Q, and midband gain. Evidently, the choice of $\alpha_1 = 0$, realizes the same function as the circuit of Fig. 1(a) and this reduces to the circuit proposed by Rao and Srinivasan earlier [2] (shown in Fig. 1(b)). The presence of finite nonzero α_1 makes the realized $(\omega_p \cdot Q)$ product of the circuit of Fig. 1(a) less than that of the Rao and Srinivasan (RS) circuit:

$$\omega_p \cdot Q = B \Bigg/ \left(1 + \alpha_1 + \frac{\alpha_2}{\alpha_3}(1 + \alpha_1 + \alpha_3)\right) \tag{1}$$

with α_1 finite and nonzero, and

$$(\omega_p \cdot Q) = B \Bigg/ \left(1 + \frac{\alpha_2}{\alpha_3}(1 + \alpha_3)\right) \tag{2}$$

with $\alpha_1 = 0$, which is more than that of (1).

Nandi recently described another bandpass filter [1], which is shown in Fig. 1(c). It is to be noted that a simple T to π transformation of resistors R_1, R_2, and R_3 yields the RS circuit with an additional parasitic resistor R_3', which does not affect the transfer function at all. For the RS circuit, the choice of suitable value of β will yield a bandpass or low-pass transfer function. For $\beta = 0$, a simple low-pass filter results.

The analysis of RS circuit, considering the open-loop cutoff frequency ω_a to be very small, yields $(\omega_p \cdot Q)$ value given by (2). However this is true for a large pole frequency only, i.e., $\omega_p >> \omega_a$. Taking into account, the finite ω_a, of the operational amplifier, for the circuit of Fig. 1(b), the pole Q can be shown to be

$$Q = \frac{\omega_p \cdot [\omega_a(R_2R_3 + R_3R_1 + R_1R_2) + BR_1R_2]}{(\omega_a^2 + \omega_p^2)(R_2R_3 + R_3R_1 + R_1R_2) + \omega_a BR_1R_2}. \tag{3}$$

Differentiating Q w·r·t·ω_p, it can be seen that Q has a maximum value of $Q_{\max}$ at a pole frequency $\omega_{p(Q_{\max})}$ where

$$Q_{\max} = \frac{1}{2} \cdot \sqrt{1 + \frac{BR_1R_2}{\omega_a(R_1R_2 + R_2R_3 + R_3R_1)}} \tag{4}$$

and

$$\omega_{p(Q_{\max})} = \sqrt{\omega_a^2 + \frac{\omega_a BR_1R_2}{R_1R_2 + R_2R_3 + R_3R_1}}. \tag{5}$$

The maximum possible value of $Q_{\max}$ is obtained when R_2 is infinite and $R_1 >> R_3$, and is given as

$$(Q_{\max})_{\max} = \frac{1}{2} \cdot \sqrt{1 + A_o} \tag{6}$$

at a pole frequency of

$$\omega_{p(Q_{\max})} = \frac{B}{\sqrt{A_o}} \tag{7}$$

where $B = A_o \cdot \omega_a$, with A_o the dc open-loop gain. For a typical operational amplifier type μA 741 with $B = 10^6 \times 2\pi$ rad/s, and $\omega_a = 25 \times 2\pi$ rad/s, the Q max obtainable is 100 at a pole frequency of 5 kHz, with ω_a equaling zero, viz. Q of 200.

Thus the finite ω_a reduces the pole Q obtainable and the gain available at resonance also. This is a property of other bandpass filters using amplifier pole and input capacitor recently described by Soliman and Fawzy [3] and Aatre–Mitra [4]. All these circuits realize a maximum $(\omega_p \cdot Q)$ product of B only. Bandpass filters with $(\omega_p \cdot Q)$ greater than B and with $(\omega_p \cdot Q)$ even infinite can be realized using the conventional differentiators with positive feedback shown in Fig. 1(d). This circuit has been derived from the Aatre–Mitra circuit. Complete discussion on its applicability to realize oscillators and bandpass filters has been described elsewhere by this author [5], [6].

It is relevant to compare the two classes of bandpass filters using the amplifier pole, using grounded and input capacitors, considered above, regarding a) compensation of the effects of finite ω_a and realizability of infinite $\omega_p \cdot Q$, b) effect of source resistance, c) noise performance, and d) the dynamic range.

a) For the input-capacitor based circuits [6], and for certain low-pass filters [7], it can be shown that a finite positive feedback of value $\beta = 1/A_o$ (where A_o is the dc open-loop gain and $B = A_o \cdot \omega_a$) will nullify the effect of finite ω_a. Then the Q max realizable for the active filters using the configuration of Figs. 1(b) and (d) is (B/ω_p). Accordingly, the gain available at the pole frequency is also increased. For values of β larger than $1/A_o$, these input capacitor based filters retain their bandpass transfer function characteristics and realize large $\omega_p \cdot Q$ (pole frequency and Q product) even infinite i.e. oscillators also [5, 6]. For RS circuit compensation of the effect of finite ω_a is possible using positive feedback through R_6 shown in dotted lines in Fig. 1(b), when

$$\frac{A_o - 1}{R_6} = \frac{1}{R_4} + \frac{1}{R_5}. \tag{8}$$

An additional condition, is required to nullify the constant term in the numerator of the transfer function, viz.

$$\frac{R_1}{R_3} = \frac{R_4}{(R_5R_6/(R_5 + R_6))}. \tag{9}$$

For realizing large $\omega_p \cdot Q$, the R_4, R_5, and R_6 values have to be changed suitably to satisfy (11) and the design equation for pole Q. For the input capacitor based circuits, e.g., Fig. 1(d), condition such as (9) is not required and simply changed β will realize the desired $(\omega_p \cdot Q)$.

b) With finite source resistance R_S, for the input capacitor based circuits (e.g., Fig. 1(d) with $\beta = 0$), the pole Q is decreased considerably when $BCR_S > 1$:

$$Q = \sqrt{BC(R_2 + R_S)}/(1 + BCR_S). \tag{10}$$

Thus for low pole frequencies, when C is large, R_S has to be extremely small. For large pole frequencies, and for large pole Q's (when $R_1 >> R_S$), the pole frequency is determined by $(R_2 + R_S)$, hence requiring that $R_S << R_2$

$$\omega_p = \sqrt{B/C(R_2 + R_S)}. \tag{11}$$

For the RS circuit, when $R_S << R_4 + R_5$, the Q value realized has no limitations such as $BCR_S > 1$, as for input capacitor based filters. Also, for large pole Q's when $R_1 >> R_S$, ω_p and Q are not affected much. With low pole Q's, since R_1 is not very large compared to R_S, for both the classes of filters, ω_p and Q are dependent on R_S.

c) The noise analysis of the circuits of Fig. 1(b) ($R_2 = \infty$) and Fig. 1(e) Soliman and Fawzy (SF) [3]), which realize identical transfer functions, using the method of Trofimenkoff *et al.* [8], will not be considered. For Fig. 1(b) the inherent noise is due to the resistors R_1, R_3, R_4, and R_5, whereas for Fig. 1(e) it is due to R_1 and R_3 only. (Various noise sources of RS circuit are shown in Fig. 1(f).) It can be shown from the noise transfer functions of the Johnson noise sources of these resistors that the inherent noise for the RS circuit is larger than that of SF circuit

$$e_{n(\mathrm{RS})}^2 = 2KT\pi BR_3(2 + Q^2) \tag{12}$$

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[1] R. Nandi, *Proc. IEEE*, vol. 67, pp. 798–799, July 1978.

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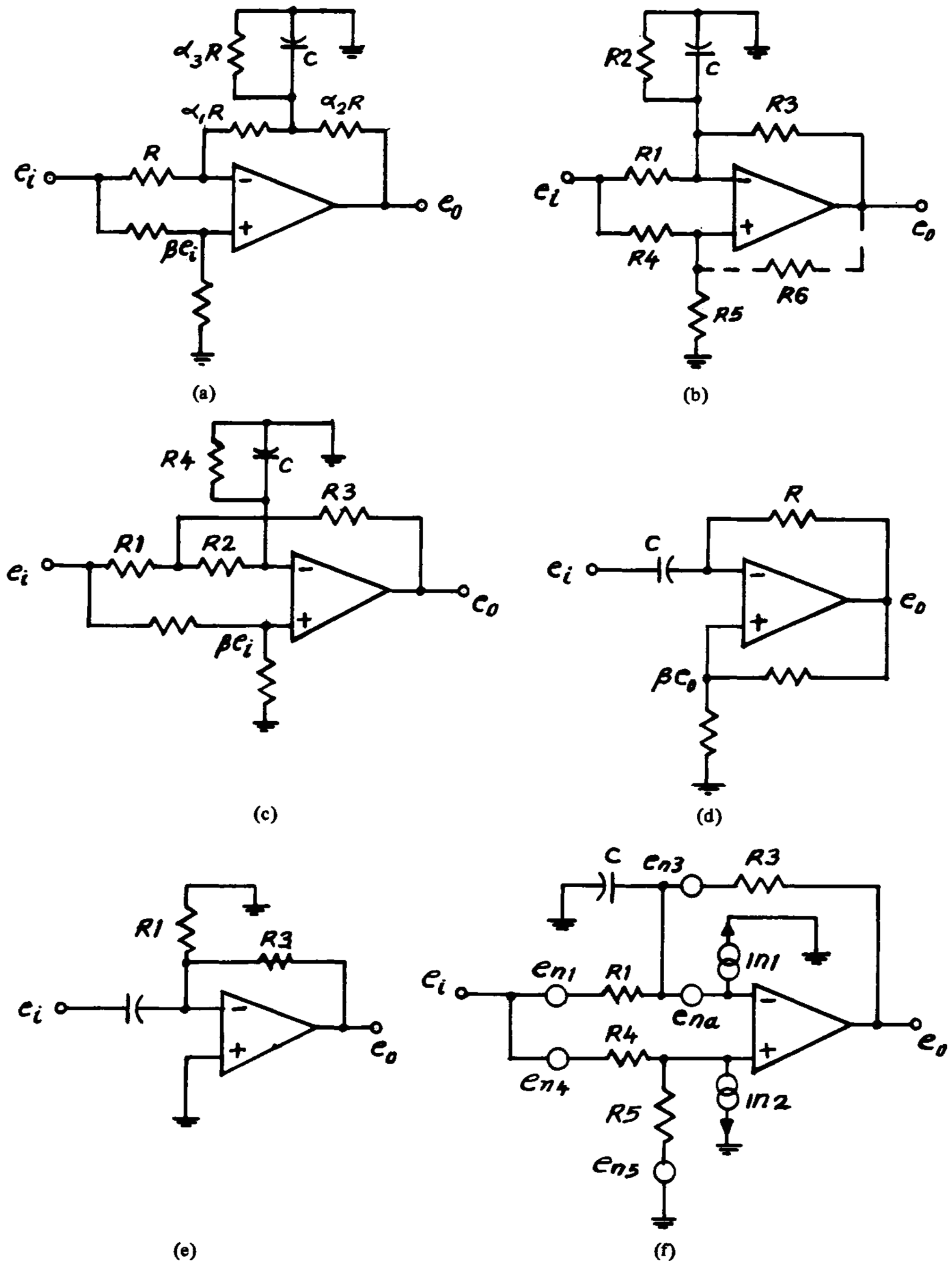

Fig. 1. Bandpass filters using the amplifier pole of Nandi (a) and (c), Rao and Srinivasan (b), the author (d), Soliman and Fawzy (e), and Rao and Srinivasan with various noise sources, (f).

and

$$e_{n(\mathrm{SF})}^2 = 2KT\pi BR_3 \cdot (2) = 4KT\pi BR_3 \tag{13}$$

where K is the Boltzmann's constant.

The RS circuit has, however, the advantage that all the resistors can be chosen to minimize the offset of the filter stage, an important requirement for cascading several stages. Under this condition,

$$\frac{R_4R_5}{R_4+R_5} = \frac{R_1R_3}{R_1+R_3}. \tag{14}$$

For ideal bandpass characteristics, RS circuit requires

$$\frac{R_5}{R_4+R_5} = \frac{R_3}{R_1+R_3}. \tag{15}$$

From (14) and (15), $R_4 = R_1$ and $R_5 = R_3$.

For realizing low offset for SF circuit, an offset resistor of value $(R_1R_3/(R_1+R_3))$ in series with the noninverting input of the OA must be provided. Under this condition, considering the inherent noise and noise sources of OA and offset resistors, it can easily be shown that *for low offset designs, RS and SF circuits are identical regarding noise performance.* For designs which do not require low offsets, the SF circuit is superior to RS circuit. These conclusions are contradictory, to what may be expected by viewing the circuits of Fig. 1(b) and (e) as integrator- and differentiator-based, respectively, and assuming that the differentiator yields more noise than integrators [10].

d) For the Rao and Srinivasan circuit, the gain at frequency (ω_p) is less than that of the Soliman and Fawzy circuit by a factor of β, and hence has more dynamic range.

Further results on bandpass filters with tunable pole frequency and constant pole Q and input capacitor based filters with prescribed midband gain ω_p and Q are described elsewhere [9].

Acknowledgment

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Reply[2] *by R. Nandi*[3]

In the course of investigations on grounded-capacitor ideal differentiators [11], the author found that the effect of the finite gain-bandwidth product of the operational amplifier (OA) could yield interesting second-order filter functions in these networks. The chief motivation f8r the design of filters and signal processing networks utilizing the dominant pole of the OA has arisen quite recently when the usual infinite-gain-mode OA circuits were shown to exhibit limited performance from mid-audio ranges for relatively high-frequency signals [12]-[14]. Subsequently, the evaluation of the effect of the finite gain-bandwidth product of the OA device was felt necessary for its utilizability at higher frequencies.

The classical differentiator network is analyzed and the effect of $\omega_a = 1/\tau_0$ (note, however, that ω_a is a low-frequency parameter usually in the range of 20 to 100 rad/s [15]) on the pole Q has also been described by Tobey *et al.* [16, p. 218]. Such an analysis may be easily extended to the Soliman-Fawzy [3] configuration, and in fact, to all the differentiators of reference [11] including the Rao-Srinivasan circuit [2] (note again that this circuit evolves from Professor Ganguly's differentiator [17]) to yield the expected conclusion that ω_a degrades pole Q at lower frequencies. Analyses based on the single-pole ideal-integrator OA model ($A = B/s$) have been shown by a number of workers to be adequately accurate and practicable [18]-[20] above mid-audio ranges and consequently, in view of the fullest utilizability of the OA device at higher frequencies, investigations on the effects of ω_a in the high-frequency range for which the networks are meant, rather than in the lower range (see Anandamohan's abstract and (7) for max Q: Fairchild μA741: B = 1 MHz, A_0 = 106 dB yield f_0 = 2.1 kHz; Signetic μA741: B = 602 kHz, A_0 = 108.9 dB [20] yield f_0 = 1.2 kHz) appear to be appropriate. Thus the basis of the comparison of high frequency performance of these networks in terms of a low-frequency parameter, although it may be of some interest to a particular designer, seems to be of limited generality.

The availability of more than one filter function capability in a network is considered to be an advantage. Rao and Srinivasan [2] have reported only the BP characteristics, and the availability of additional LP function in such grounded-capacitor differentiators [11] was not spelled out in the open literature until this author proposed a fuller description of the dual-function capability of such a circuit together with their experimental verification [21]. To this end, the author wishes to point out that the elegant circuit of Soliman and Fawzy [3] can be designed to yield a dual-function by a simple terminal adjustment of the noninverting input terminal of the OA either to the ground (BP) or to the input (LP) point. Work on such dual-function capability for all the circuits of [11] is now being carried out and would be reported some time later.

Finally, in the light of the above discussions, the author feels that the criterion of originality and novelty supported by indepth analytical and practical investigations, probably the proper and primary evaluating factors, have been satisfied in the work presented in reference [21], and the author finds hardly any point in wondering to the possibilities of topological alternatives (T to π transformation or RC:CR transformations etc.) which sometimes are nothing but very usual and likely situations that creep up in the event that a newly emerging field is being attacked by stalwarts. For example, Aatre and Mitra's [20] structure may be viewed as a more generalized version of the HO-Chiu [22] circuit that has been worth re-cited by Venkateswaran [23] for its additional band-elimination function; Soliman and Fawzy's work [3] leads to a useful tunable filter which, after omission of an appropriate resistor, gives back the classical differentiator [16]; Ahmed's [24] recent sine-wave generator is the well-known Wien's version but is derived nicely from the newer FDNR concepts, Holt *et al.* [25] and author's [26], although active-*R* inductors are quite alike, their application aspect, along with the contributions referred to above, have substantially enriched the maturity of the subject and the totality of the technical literature. The author, therefore, claims the validity of [21] as a creditable piece of work on a modern topic supported by adequate technical information which is hoped to be worthy of stimulating interests and utility to the knowledgeable readership of the Electrical Engineering profession.

[2] Manuscript received August 2, 1979.
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Realization of an All-Pass Transfer Function Using the Second Generation Current Conveyor

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***Abstract*—The second generation current conveyor (CCII) is used together with only four components to realize an all-pass transfer function.**

A number of circuits realizing all-pass transfer functions have been reported in the literature. Most of these circuits use operational ampli-

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