$$c_{n,k} = \begin{cases} \sin\left(\frac{\pi(k+0.5)n}{N} + \theta_k\right), & \text{k odd} \\ \cos\left(\frac{\pi(k+0.5)n}{N} + \theta_k\right), & \text{k even} \end{cases} \quad (6)$$

$$(-1)^m c_{n-(2m+1)N,k} = (-1)^k s_{n,k} \quad (7)$$

$$(-1)^m c_{n-2mN,k} = c_{n,k} \quad (8)$$

Using these relations, we can express $s(n)$ as

$$s(n) = \sum_{k=0}^{N-1} \sum_{m'=-\infty}^{+\infty} x'_k(m') g_k(n - m'N) \quad (9)$$

where

$$x'_k(m') = \begin{cases} (-1)^m x_k(m), & m' = 2m+1 \\ (-1)^m x_{N+k}(m), & m' = 2m \end{cases} \quad (10)$$

$$g_k(n) = g(n)\cos\left(\frac{\pi(k+0.5)n}{N} + \phi_k\right) \quad (11)$$

In the receiver, the following expressions of the demodulated signals $\hat{x}_k(m)$ can be obtained though a similar process:

$$\hat{x}'_k(m) = \sum_{n=-\infty}^{+\infty} r(n) h_k(mN - n) \quad (12)$$

$$\hat{x}_k(m) = (-1)^m \hat{x}'_k(2m+1) \quad (13)$$

$$\hat{x}_{N+k}(m) = (-1)^m \hat{x}'_k(2m) \quad (14)$$

$$h_k(n) = g(n)\cos\left(\frac{\pi(k+0.5)n}{N} + \hat{\theta}_k - ((-1)^k - 1)\frac{\pi}{4}\right) \quad (15)$$

Thus, we obtain a generalised cosine modulated filter bank structure for the O-QAM system. The filters $g_k(n)$ and $h_k(n)$ are related to the prototype $g(n)$ by cosine modulation.

Design and implementation: By letting the modulation phase factors θ_k and $\hat{\theta}_k$ all be zero, the model is the same as in [3]. If we assume that the length of $g(n)$ is $2MN$, the implementation needs two N-point type-III DCT processing and a $2N$ FIR filter operating with length M for N output sampling points. Although the implementation is more efficient compared with the DFT method in [2], the design is also difficult, when the perfect reconstruct property is required. If we select the phase factors as

$$\theta_k = -\left(M - \frac{1}{2N}\right)(k+0.5)\pi - (-1)^k\frac{\pi}{2} + \frac{\pi}{4} \quad (16)$$

$$\hat{\theta}_k = \left(M - \frac{1}{2N}\right)(k+0.5)\pi - (-1)^k\frac{\pi}{2} + \frac{\pi}{4} \quad (17)$$

then the filters $g_k(n)$ and $h_k(n)$ are the same as the synthesis and analysis filters in [4], respectively. The excellent design technique using the two-channel lossless lattice [4] can be directly employed based on the simple complementary relation between transmultiplexers and sub-band analysis/synthesis systems. In the implementation, only one N-point type-IV DCT processing and the operation of N two-channel lossless lattices, each with M sections, are required for N output sampling points.

Conclusions: The proposed structure extends the sub-band filter bank model of the O-QAM system to a more general form. On the basis of this stucture, the design and implementation of the O-QAM system can be performed very efficiently by selecting proper modulation phase factors.

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High-frequency linear transformation elliptic filters employing minimum number of OTAs

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Indexing terms: Active filters, Elliptic filters, Operational transconductance amplifiers

High-frequency linear transformation (LT) elliptic filters employing a minimum number of operational transconductance amplifiers (OTAs) and capacitors are presented. Based on the LT, an OTA-C elliptic filter can be realised efficiently by adding capacitors to the relative all-pole filter as passive prototype counterparts. The canonical filter with all nodes having a desired capacitance, which is suitable for high-frequency operation, is developed. As an example, a new third-order elliptic lowpass filter is synthesised. Experimental results are obtained to verify the theoretical analysis. Furthermore, the proposed circuit can be extended to higher-order filters.

Introduction: The advantages of doubly terminated LC ladder filters are that they have very low sensitivities in the passband and low component spread. However, most of the proposed OTA-C synthesis methods [1 – 4] for emulating high-order LC ladder filters, especially elliptic filters, are subject to either complicated design procedures or extra numbers of OTAs. Moreover, most of them are unsuitable for high-frequency operation because of parasitic capacitance effects [5]. LT [6, 7] filters possess the characteristics that one can realise every section of the original ladder prototype using active elements individually. In this Letter, we use the LT to design OTA-C elliptic filters by adding capacitors to all-pole filters as passive prototype counterparts. Moreover, our proposed canonical filter has not only a minimum number of OTAs but also good high-frequency characteristics.

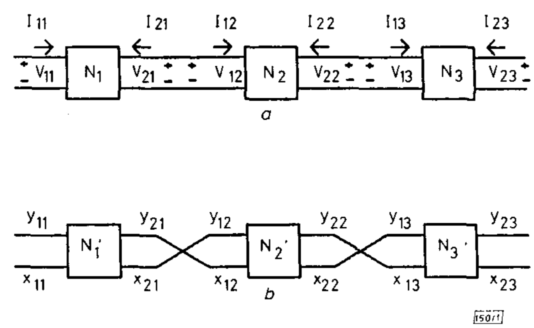

Fig. 1 *Original and transformed two-port networks*

a Original two-port networks for $n = 3$
b Transformed two-port networks

Design principle and procedure: Using the principle of LT, we can divide the LC filter prototype into several two-port networks, as shown in Fig. 1*a*, and transform the input/output voltage and current variables of a two-port network into two new variables x_{ji}

and y_{ji}. Their characteristics can be given as

$$\begin{bmatrix} x_{ji} \\ y_{ji} \end{bmatrix} = S_{ji} \begin{bmatrix} V_{ji} \\ I_{ji} \end{bmatrix} = \begin{bmatrix} \alpha_{ji} & \beta_{ji} \\ \gamma_{ji} & \delta_{ji} \end{bmatrix} \begin{bmatrix} V_{ji} \\ I_{ji} \end{bmatrix} \quad (1)$$
$$j = 1, 2 \text{ and } i = 1 \text{ to } n$$

where S_{ji} is the transformation matrix, n is the order of the filter, α_{ji} and γ_{ji} are dimensionless, and the units of β_{ji} and δ_{ji} are ohms, so that the dimensions of x_{ji} and y_{ji} are volts. If x_{ji} represents the inputs and y_{ji} represents the outputs, we can describe a new two-port network as shown Fig. 1*b*. To avoid the complexity of general interconnection between two new two-ports, the cross-cascade interconnection [6] is introduced and is also shown in Fig. 1*b*. As an example, for a third-order filter, if we choose

$$\begin{bmatrix} x_{21} \\ y_{21} \end{bmatrix} = \begin{bmatrix} 0 & -R \\ 1 & 0 \end{bmatrix} \begin{bmatrix} V_{21} \\ I_{21} \end{bmatrix} \quad (2a)$$

$$\begin{bmatrix} x_{12} \\ y_{12} \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & R \end{bmatrix} \begin{bmatrix} V_{12} \\ I_{12} \end{bmatrix} \quad \begin{bmatrix} x_{22} \\ y_{22} \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & -R \end{bmatrix} \begin{bmatrix} V_{22} \\ I_{22} \end{bmatrix} \quad (2b)$$

$$\begin{bmatrix} x_{13} \\ y_{13} \end{bmatrix} = \begin{bmatrix} 0 & R \\ 1 & 0 \end{bmatrix} \begin{bmatrix} V_{13} \\ I_{13} \end{bmatrix} \quad (2c)$$

it can be seen that $y_{21} = V_{21} = V_1 = V_{12} = x_{12}$ and $x_{22} = V_{22} = V_3 = V_{13} = y_{13}$. Because transformed node voltages are equal to untransformed counterparts, we can separate the floating capacitors in the series arm of elliptic filters from the passive prototype, and consider all-pole filters only.

According to the proposed design method, an LT elliptic filter can be synthesised by the following procedures:

(i) Separate the floating capacitors in the series arm of elliptic filters from the passive prototype, and consider all-pole filters only.

(ii) Divide the all-pole ladder prototype into several sections and choose appropriate transformation matrices to carry out their x-y domain transfer fuctions.

(iii) Replace the transfer functions with their corresponding OTA-C circuits and connect neighbouring sections with the cross-cascade interconnection.

(iv) Add the previous separated floating capacitors to the synthesised all-pole OTA-C filter at the same nodes.

(v) Determine the values of transconductances and capacitances via the derived design equations.

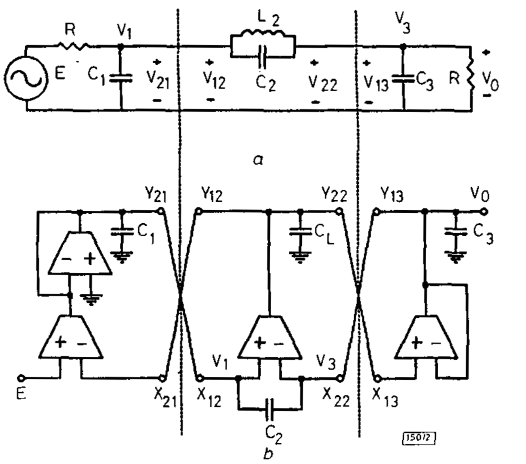

Fig. 2 *Third-order elliptic LC lowpass filter prototype and corresponding OTA-C filter of Fig. 2a*

a Third-order elliptic *LC* lowpass filter prototype
b Corresponding OTA-C filter of Fig. 2*a*

Design example and experimental result: To demonstrate the efficiency of the proposed design method, a third-order *LC* elliptic lowpass filter is realised. Fig. 2*a* shows its prototype with a 1dB ripple and 50kHz bandwidth. At first we can separate the floating capacitor C_2 from the passive prototype, divide the filter prototype into three sections, and choose transformation matrices in eqn. 2. Using LT, the transfer functions can be respectively derived from eqn. 2 as

$$y_{21} = \frac{E - x_{21}}{sRC_1 + 1} = \frac{E - x_{21}}{s\frac{C_1}{g_m} + 1} \quad (3a)$$

$$y_{12} = y_{22} = \frac{x_{12} - x_{22}}{s\frac{L_2}{R}} = \frac{x_{12} - x_{22}}{s\frac{C_L}{g_m}} \quad (3b)$$

$$y_{13} = \frac{x_{13}}{sRC_3 + 1} = \frac{x_{13}}{s\frac{C_3}{g_m} + 1} = V_0 \quad (3c)$$

We replace eqn. 3 with the corresponding OTA-C circuits and join them together with the crosscascade interconnection. Finally, we add C_2 to the synthesised circuit at nodes V_1 and V_3. Fig. 2*b* shows the completed filter, and it employs only four identical OTAs. The overall transfer functions of the *LC* prototype and proposed circuit are the same and yield

$$\frac{V_0}{E} = \frac{s^2 L_2 C_2 + 1}{\left\{ s^3 L_2 R (C_1C_2 + C_2C_3 + C_3C_1) + s^2 L_2 (C_1 + 2C_2 + C_3) + s\left(RC_1 + \frac{L_2}{R} + RC_3\right) + 2 \right\}}$$
$$= \frac{s^2 \frac{C_L C_2}{g_m^2} + 1}{\left\{ s^3 \frac{C_L}{g_m^3} (C_1C_2 + C_2C_3 + C_3C_1) + s^2 \frac{C_L}{g_m^2} (C_1 + 2C_2 + C_3) + s\frac{1}{g_m}(C_1 + C_L + C_3) + 2 \right\}} \quad (4)$$

This canonical filter is suitable for high-frequency operation, because all nodes have a desired capacitance and the parasitic capacitances can be incorporated in the circuit capacitors [8]. The required number of OTAs in Fig. 2*b* is less than that of the previous works [1 – 4], and a comparison is shown in Table 1. Fig. 2*b* is verified experimentally by using the bipolar OTAs LM13600 along with discrete capacitors $C_{a1} = C_{a3} = 20$nF, $C_{a2} = 2.63$nF and $C_b = 10$nF. Its amplitude response is shown in Fig. 3. Experimental results confirm the theoretical analysis.

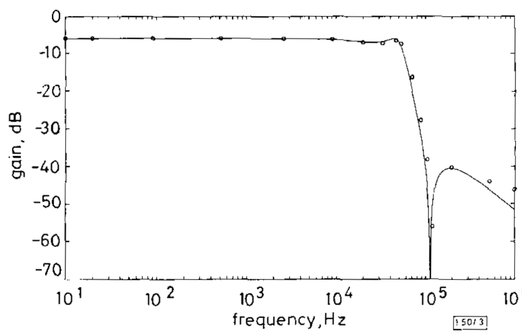

Fig. 3 *Amplitude response of Fig. 2b*

——— theoretical
O experimental

Table 1: Comparison between proposed filter and previous works

Authors	No. of OTAs	No. of grounded capacitors	No. of floating capacitors	High-frequency operation
A.C.M. de Queiroz [1]	5	3	1	yes
L.P. Caloba [2]	7	3	1	yes
D.Y. Kim [3]	6	3	1	no
M.A. Tan	5	3	1	no
Proposed filter	4	3	1	yes

Conclusions: Systematic and effective design methods and procedures for realising high-order LT elliptic OTA-C filters are presented. The proposed filter has a minimum number of OTAs and is suitable for high-frequency operation. Furthermore, it can be extended to higher-order filters. It is a useful tool for the design of high-frequency filters.

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Transition based LMS adaptive filter

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Indexing terms: Adaptive filters, Cochannel interference

The performance of conventional adaptive filters used for cochannel interference suppression degrades when the desired signal and the interference have similar characteristics. The author proposes a technique that makes use of the different data clock phases of the desired signal and the interference to separate the data transitions of the desired signal from those of the interference. This principle is used to modify a least mean square adaptive filter. Simulation results and an expression for the weight error vector are presented.

Introduction: Adaptive filtering has been widely used to remove cochannel interference that is not correlated [1]. The performance of these adaptive filter schemes degrades when the signal not of interest (SNOI) to be removed has characteristics simliar to the signal of interest (SOI). This Letter proposes a modification to the existing least mean square (LMS) transversal adaptive filter to enable it to separate similar cochannel transmissions, when the modulation used is linear. The proposed technique can also be extended to other algorithms.

Cochannel interference problem: When the input to the LMS adaptive filter consists of two or more cochannel signals, whose second-order statistics are similar, which is the case if the transmissions have the same carrier frequencies, same data rates and similar data distributions, it is not possible to separate the SOI from the SNOI as the weight vector would now be biased by an interference component. This is easily shown as follows: The received signal can be written as

$$\mathbf{Y} = \mathbf{S} + \sum_{j=1}^{M} \mathbf{I}_j + N \tag{1}$$

where $\mathbf{S}$ is the transmitted signal, $\mathbf{I}_j$ is the jth cochannel interference, M is the number of simultaneous cochannel transmissions, and N is Gaussian noise with variance σ_n^2. The conventional LMS adaptive filter update for the solution to the normal equations is given by

$$\mathbf{W}(K+1) = \mathbf{W}(K) + \mu\left(d(K) - \mathbf{W}^T(K)\mathbf{Y}(K)\right)\mathbf{Y}(K) \tag{2}$$

where $d(K)$ is the desired signal at time K. Defining the weight error vector $\mathbf{V}(K) = \mathbf{W}(K) - \mathbf{W}^*$, where $\mathbf{W}^*$ is the optimal weight vector, we can express eqn. 2 as

$$\mathbf{V}(K+1) = \mathbf{V}(K) + \mu\left(n(K) - \mathbf{V}^T(K)\mathbf{Y}(K)\right)\mathbf{Y}(K) \tag{3}$$

The weight error power [2], defined as $trE[\mathbf{V}\mathbf{V}^T]$ converges to

$$\lim_{K\to\infty} trE[\mathbf{V}(K)\mathbf{V}^T(K)] = \frac{\mu_{LMS} L \sigma_n^2 E[y^2]}{2E[y^2] - \mu_{LMS}\{(L-1)E[y^2]E[y^2] + E[y^4]\}} \tag{4}$$

where L is the order of the filter. The effect of the interference can be studied by looking at $E[y^2]$ and $E[y^4]$. The fourth-order statistics can be approximated by second-order statistics since the input can be considered to be Gaussian. Neglecting the crossterms involving Gaussian noise, $E[y^2]$ can be written as

$$E[\mathbf{Y}(K)\mathbf{Y}^T(K)] = \left(E[SS^T] + E[n^2] + 2E\left[\left(\sum_{j=1}^{K}\mathbf{I}_j\right)\mathbf{S}^T\right] + E\left[\sum_{j=1}^{K}\mathbf{I}_j\mathbf{I}_j^T\right] + E\left[\sum_{i=1}^{K}\mathbf{I}_i \sum_{j=1, j\neq i}^{K}\mathbf{I}_j^T\right]\right) \tag{5}$$

The main interference component is $E[\Sigma_{j=1}^{K} \mathbf{I}_j\mathbf{I}_j^T]$ which reflects the correlation of the cochannel interferers with themselves. This would be a strong component especially when the interferers are much stronger than the SOI. In addition, the third and fifth terms in eqn. 5, although considered to be zero under the assumption that the different transmissions are independent, reflect a time-varying interference component dependent on the length of the filter and the data transmitted by the different sources. The effect of neglecting all these components is that the resulting solution is not optimal.

Transition based LMS adaptive filter: One feature that can be used to separate the SOI from the interference is the difference in the phase of the data clocks used in all the transmissions concerned. Using the phase information of the data clock would enable us to determine the transitions in the received data stream that correspond to those of the SOI only. Since the transitions in the baseband signal completely characterise the digital data transmitted, it would then be possible to use this scheme to achieve cochannel interference suppression. Assuming that data timing is available at the receiver and that the demodulated data are oversampled, the transitions in the input signal can be determined and the desired transitions selected by having the weight update of the TB-LMS filter expressed as

$$\mathbf{W}(K+1) = \mathbf{W}(K) + \mu\left(d(K) - \mathbf{W}^T(K) f(\mathbf{Y}(K))\right) f(\mathbf{Y}(K)) \tag{6}$$

where $f(\mathbf{Y}_K)$, assuming a simple delay and subtract transition detector, can be written as

$$f(\mathbf{Y}(K)) = (\mathbf{Y}(K) - \mathbf{Y}(K-1)) \times Win(K) \tag{7}$$

$$Win(K) = 1, \text{ leading or lagging edge of the date clock of the SOI} \tag{8}$$

Weight error vector: The expression for the weight error power for the TB-LMS filter is