

the gate width range, devices of 2-3 mm gate width have delivered (in class A) power densities of 0.45 W/mm with associated gains (-1 dB) of 13 dB and power added efficiencies of 36%. Figs. 4 and 5 show the power transfer and the power-added efficiency characteristics for representative devices having total gate widths of 2.46 and 14.76 mm, respectively.

We believe that the true performance of the 14.76 mm device is better than that depicted by Fig. 5, since the data were acquired while the device was only partially matched on the input side.

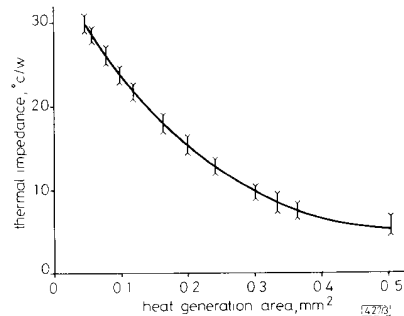


Fig. 3 Measured thermal impedance against heat generation area

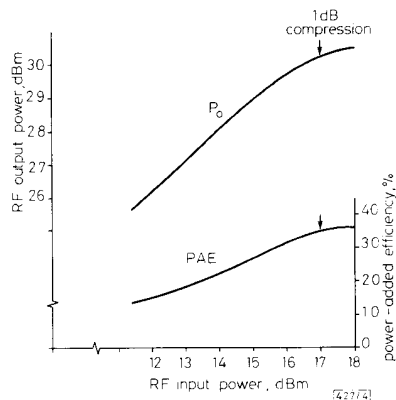


Fig. 4 Power transfer and power-added efficiency characteristics

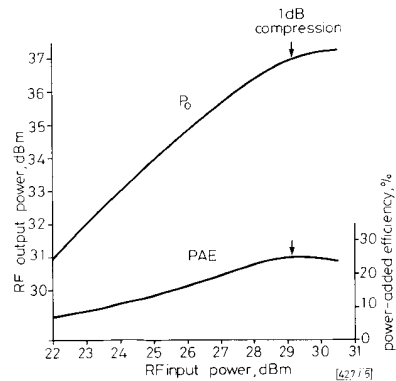


Fig. 5 Power transfer and power-added efficiency characteristics

Conclusions: A GaAs foundry process has been adapted to enable the construction of elemental devices for future power GaAs MMICs. Novel integral heatsink structures have been developed, which involve etching 'bath-tub' shaped cavities under the FET devices, filling those bath-tubs with Au, and finally re-planarising the back face of the GaAs wafer. The

efficacy of the integral heat sinks has been verified by thermal simulations, measurements of thermal impedance and sectioning of devices.

A range of power FET devices, having a front-face layout fully compatible with a GaAs 'foundry' IC process, have been characterised at 3 GHz. The devices has shown excellent performance in terms of gain, peripheral power density, and power added efficiency.

A route for fabricating power GaAs MMICs on a manageable substrate thickness has thus been demonstrated.

Acknowledgments: The authors would like to acknowledge the contributions of the following persons: Dr. R. W. W. Charlton and colleagues in the GaAs IC processing facility at Plessey Caswell, for front-face processing of GaAs IC wafers; Mr. B. P. Davies, for contributing to the development of the ion-implant profile; Mr. J. A. Sparrow and colleagues for chip mounting, bonding, and device sectioning; and finally Dr. P. W. Webb of the University of Birmingham, UK for performing thermal simulations. This work was funded in part by Siemens Plessey Radar Ltd.

P. R. BESTWICK
I. DAVIES
K. C. VANNER

23rd May 1990

Plessey Research Caswell Ltd
Caswell, Towcester, Northants.
NN12 8EQ, United Kingdom

Reference

- SUFFOLK, J. R., *et al.*: 'Yield enhancement techniques for GaAs MMICs'. Digest of 1989 US Conf. GaAs Manufacturing Technology, San Diego, October 24-26 1989, pp. 28-32

PROGRAMMABLE MULTIPLEXED SWITCHED-CAPACITOR FILTERS

Indexing terms: Filters, Switched capacitor filters, Multiplexers and multiplexing

An area-efficient design approach for implementing a fully programmable N th order switched-capacitor filter is proposed. Since programmable capacitor arrays (PCAs) consume vast amounts of area the proposed approach makes use of time-multiplexing resulting in the need for only three PCAs.

Introduction: With switched-capacitor (SC) filters, multiplexing (or time-sharing) has traditionally been used to reduce the number of operational-amplifiers.¹⁻³ We choose to multiplex the input-summing capacitors of the integrators resulting in an area efficient method to realise N th order fully programmable SC filters. The final design requires three programmable capacitor arrays (PCAs)* together with $2N$ clock phases.

It should be mentioned here that without the use of multiplexing, the minimum number of PCAs required would be $2N$. It is typical to use at least $3N$ PCAs to maintain reasonable filter performance with respect to sensitivity and dynamic range. This requirement for a large number of PCAs is the major reason that high-order fully programmable SC filters are not readily available. The drawback of multiplexing is the large number of clock phases required and the shorter time available for charging and discharging capacitors.

To realise arbitrary transfer-functions the usual approach is to make use of cascades of biquads. The difficulty with this approach is that the optimum biquad structure depends

* A PCA is an array of capacitors which allows a user to select a capacitor value using digital control.

strongly on the transfer-function being realised such as bandpass, high-pass, notch, etc. and on the value of the pole-Q. An alternative approach is to make use of the orthonormal ladder filter structure⁴ where a single structure is used for arbitrary transfer-functions and is inherently scaled for optimum dynamic range. The proposed design makes use of an exact design technique using lossless discrete integrators (LDI) in the orthonormal ladder filter structure resulting in a parasitic insensitive design and no delay-free loops—a necessary requirement to make use of multiplexing. The same filters appear to have sensitivity and dynamic range performances comparable to an optimum design of a cascade of biquads.

Design: As an example, consider a third order LDI-based orthonormal SC filter as shown in Fig. 1. The set of equations which describes this filter is given by

$$\begin{bmatrix} X_1(z) \\ X_2(z) \\ X_3(z) \end{bmatrix} = z^{-1} \begin{bmatrix} X_1(z) \\ X_2(z) \\ X_3(z) \end{bmatrix} + \begin{bmatrix} 0 & a_{12} & 0 \\ z^{-1}a_{21} & 0 & z^{-1}a_{23} \\ 0 & a_{32} & a_{33} \end{bmatrix} \times \begin{bmatrix} X_1(z) \\ X_2(z) \\ X_3 \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \\ b_3 \end{bmatrix} U(z) \quad (1)$$

$$Y(z) = [c_1 \quad c_2 \quad c_3] \begin{bmatrix} X_1(z) \\ X_2(z) \\ X_3(z) \end{bmatrix} + dU(z) \quad (2)$$

To use this structure efficiently, a multiplexing scheme is introduced where each of the three PCAs are time-shared by

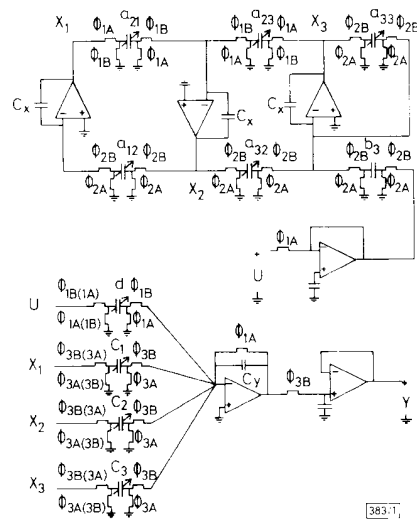


Fig. 1 Third order programmable multiplexed switched capacitor filters

three coefficients. These are three intervals of non-overlapping clock phases, as shown in Fig. 2. During each interval, three coefficients are serviced† by the PCAs, for e.g., during the first interval, coefficients a_{21} , a_{23} and d are serviced. Coefficients are grouped in threes as; (a_{21}, a_{23}, d) , (a_{12}, a_{32}, a_{33}) and (c_1, c_2, c_3) . Two different clocking schemes are given for d and c coefficients. The clocking scheme in parenthesis is used for positive coefficient values. The scheme without parenthesis is used for negative ones. The coefficients are to be serviced in such a way that the even ordered states which use the past states are computed before the odd ordered states (eqn. 1). Once all the states are computed, a weighted sum of all the states as well as the input is obtained. During the first clock

† The term service is used to illustrate the action of substituting for the feed-in capacitors in the switched-capacitor filter by PCAs during an interval.

phase ϕ_{1A} the output summer is cleaned. A new sum is calculated in each clock period. In general, $2N$ clock phases and three PCAs are required to realise an N th order filter.

Experimental results: A discrete prototype of a fifth-order filter was designed and built using the approach described above. Eight-bit binary PCAs were used with the value of the

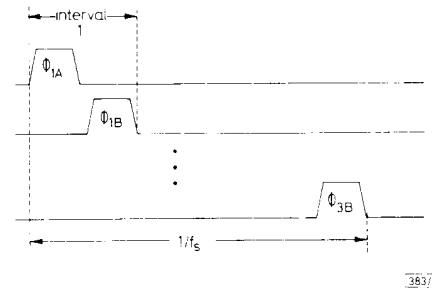


Fig. 2 Clocks used for multiplexing PCAs

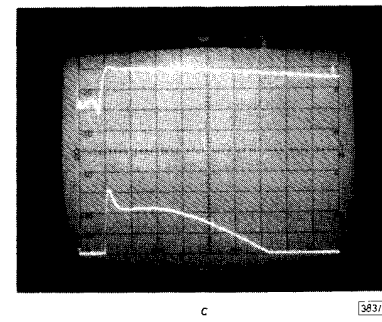
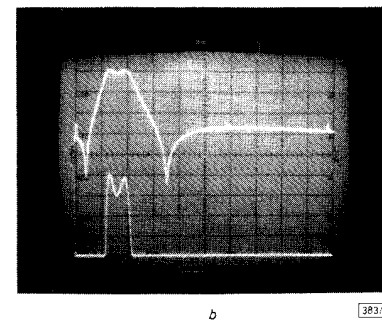
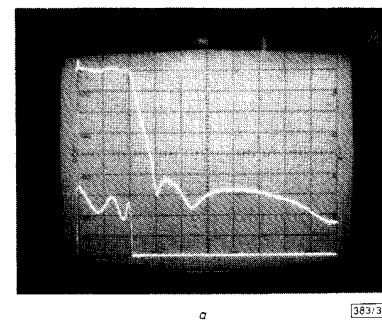


Fig. 3 Transfer functions obtained from discrete prototype
 a 5th order equiripple lowpass
 b 4th order bandpass
 c 3rd order highpass

PCA controlled by EPROMs. The input sampling rate of the filter was set to 10 kHz although internally the on-time of each clock phase was just under 10 μ s. To demonstrate the usefulness of this filter, the coefficients for different transfer-functions were programmed into the EPROMs. Fig. 3 shows the frequency response for three of these transfer-functions. Note in Fig. 3a a transmission zero is shown at half the sampling rate since an exact design technique was used although the design is LDI-based. Also note in Fig. 3 that the passband droop is due to the sample-and-hold inherent in the filter. Other filters that have been programmed are an all-pass and a Chebyshev type.

Conclusion: An area efficient method of designing programmable SC filters has been presented. It was shown that multiplexing could be used to reduce the number of PCAs to only three for an N th order filter. A discrete prototype filter was constructed and experimental results given, demonstrating the practicality of the approach.

Acknowledgments: The authors would like to thank Professor W. M. Snelgrove for aid in developing the LDI-based orthonormal structure.

X. F. WANJA
D. A. JOHNS
A. S. SEDRA

16th May 1990

Department of Electrical Engineering
University of Toronto
Toronto, Ontario, Canada, M5S 1A4

References

- ROBERTS, G. W., SNELGROVE, W. M., and SEDRA, A. S.: 'Switched-capacitor realisation of N th-order transfer-function using a single multiplexed op-amp', *IEEE Trans.*, 1987, **CAS-34**, pp. 140-148
- MONTECCHI, F.: 'Time-shared switched-capacitor ladder filters insensitive to parasitic effects', *IEEE Trans.*, 1984, **CAS-31**, pp. 349-353
- LAKER, K. R., FLEISCHER, P. E., and GANESAN, A.: 'Parasitic insensitive biphase switched capacitor filters realized with one operational amplifier per pole pair', *Bell Syst. Tech. J.*, 1982, **61**, (5), pp. 685-707
- JOHNS, D. A., SNELGROVE, W. M., and SEDRA, A. S.: 'Orthonormal ladder filters', *IEEE Trans.*, 1989, **CAS-36**, pp. 337-343

PROPAGATION MEASUREMENTS AT 1.7 GHz FOR MICROCELLULAR URBAN COMMUNICATIONS

Indexing terms: Radio links, Attenuation, Mobile radio systems

Narrowband and wideband propagation results in and around a building are reported, together with their implications on the choice of data rates and the possible need for equalisation in any new mobile communication system working at 1.7 GHz.

Introduction: Radio propagation measurements have always been recognised as having an important part to play in the design of complex communication systems. This is increasingly true in areas where multipath propagation can cause significant distortion in transmitted signals, such as within buildings and urban areas. With the advent of GSM in Western Europe, linear modulation in North America and the recent announcements of proposals for PCN-type systems in the UK, future communications are moving higher in frequency, where the problems of multipath propagation and signal attenuation become more acute.

This letter presents the results of narrowband and wideband propagation measurements that have been conducted at Bristol, in and around buildings, at 1.7 GHz. The narrowband measurements were devised to measure attenuation of radio

signals within buildings, and the wideband techniques to measure multipath impulse responses and their associated RMS delay spreads.

Narrowband measurements: For the narrowband measurements, it was assumed that the attenuation in a signal path at 1.7 GHz could be represented by¹

$$A_t = L + F_L + 10n \log_{10}(d) \quad (1)$$

where A_t represents the total attenuation, F_L the loss through a floor, $10n \log_{10}(d)$ the path loss and L any other losses or gains, which will include factors such as the gains of the antennas used and losses through walls. L can be represented by the power loss at 1 m.²

The data for the 1.7 GHz attenuation measurements were collected on a raster-scan technique, where the transmitter was left stationary and the receiver scanned over regular-sized areas spaced at 2 m intervals.

All the results were collected from within Queen's Building, which houses the Engineering Faculty at Bristol University. The building is relatively modern (opened in the late 1950s) and is based on prestressed concrete and brickwork with walls up to 40 cm thick. The attenuation measurements were conducted in two corridors, one above the other, and the rooms that adjoined them.

The results for the measurements in the corridors can be seen in Fig. 1a. Three useful values that can be extracted by linear regression are the path loss index, n , the floor loss, F_L , and the losses due to antennae gain and other obstructions, L , which are defined in eqn. 1. The figures shown in brackets are those measured at 815 MHz by Arnold *et al.*²

For the results on the same floor

$$n = 1.495 \quad F_L = 00.000 \text{ dB} \quad L = 21.189 \text{ dB} \\ (n < 2 \quad F_L = 00.000 \text{ dB})$$

For the cross-floor results

$$n = 1.524 \quad F_L = 27.176 \text{ dB} \quad L = 21.189 \text{ dB} \\ (n < 2 \quad F_L = 26 \text{ dB for a metal-partitioned floor})$$

The results for the measurements in the rooms can be seen in Fig. 1b.

For the results on the same floor

$$n = 3.250 \quad F_L = 00.000 \text{ dB} \quad L = 23.638 \text{ dB} \\ (n = 3.62 \quad F_L = 00.000 \text{ dB})$$

For the cross-floor results

$$n = 3.311 \quad F_L = 12.742 \text{ dB} \quad L = 23.638 \text{ dB} \\ (n = 3.93 \quad F_L \approx 10 \text{ dB})$$

The interesting conclusions to be drawn from these results are, first, that the index of path loss for the corridor measurements is less than that of free space. This phenomenon has been observed at other frequencies,^{2,3} and has always been attributed to a channelling effect. Secondly, the increase in the index of path loss when moving from the corridor to within the rooms adjoining it; the received signal now has no line of sight, and so the channelling effect noted for the corridor measurements cannot occur. Thirdly, the increase in L when moving into the rooms. This can be explained by the attenuation due to the doors of the room (2.499 dB). Finally, the floor loss decreased from 27.176 dB to 12.472 dB when moving into the rooms. The reason for this difference lies in the construction of the building. The ceiling of the corridor is much thicker than that of the rooms, and consequently the attenuation is much higher.

Wideband measurements: One of the possible scenarios for future mobile communications is a high-data-rate TDMA digital cellular network.⁴ Such systems have been suggested before for frequencies in the millimetre-wave regions.^{5,6} One of the main problems is echoes in transmission caused by multipath propagation. The RMS delay spread has been