References

[1] J. Cohen and M. Gilden, "Temperature stability of an MIC Gunn-effect oscillator," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-21, pp. 115–116, Feb. 1973.

[2] H. Abe, Y. Takayama, A. Higashisaka, and H. Takamizawa, "A highly stabilized low-noise GaAs FET integrated oscillator with a dielectric resonator in the *C*-band," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-26, pp. 156–162, Mar. 1978.

[3] T. Yamaguchi, Y. Komatsu, T. Otobe, and Y. Murakami, "Newly developed ternary (Ca, Sr, Ba) zirconate ceramic system for microwave resonators," *Ferroelectrics*, vol. 27, pp. 273–276, 1980.

[4] W. R. Day. Jr., "Dielectric resonators as microstrip-circuit elements," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-18, pp. 1175–1176, Dec. 1970.

[5] T. G. Iveland, "Dielectric resonator filter for application in microwave integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-19, pp. 643–652, July 1971.

[6] M. W. Pospieszalski, "Cylindrical dielectric resonators and their applications in TEM line microwave circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, pp. 233–238, Mar. 1979.

[7] Y. Garault and P. Guillon, "Coupling coefficient between a microstrip transmission line and a dielectric resonator disk for application to bandpass filters," in *Proc. 5th Colloquium on Microwave Communication* (vol. IV, *Microwave Circuits and Devices MT-19*), June 1974, pp. 115–164.

[8] S. B. Cohn, "Microwave bandpass filters containing high-Q dielectric resonators," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-16, pp. 218–227, Apr. 1968.

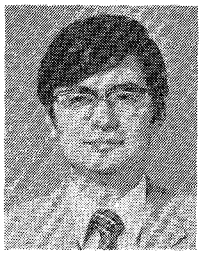
[9] Korn and Korn, *Mathematical Handbook for Scientists and Engineers*. New York: McGraw-Hill, 1961.

[10] E. L. Ginston, *Microwave Measurements*. New York: McGraw-Hill, 1957.

[11] D. S. Jones, *The Theory of Electromagnetism*. New York: Pergamon, 1964, pp. 17–19.

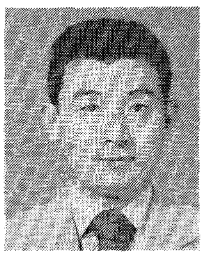
[12] W. Schlosser and H. G. Unger, "Partially filled waveguide and surface waveguide of rectangular cross section," in *Advances in Microwave*. New York: Academic Press, 1966, vol. 1, pp. 319–387.

[13] P. Skalicky, "Coupling coefficient between dielectric resonator and microstrip line," *Electron. Lett.*, vol. 17, pp. 99–100, Jan. 1981.



Yasutoshi Komatsu was born in Tokyo, Japan, in 1946. He received the B.S. degree in electrical engineering from University of Tokyo, Japan, in 1969.

In 1969, he joined the Sony Corporation Research Center, Yokohama, Japan. From 1969 to 1975, he worked on the development of crystal filters with high coupling resonators and on surface acoustic wave (SAW) devices. He developed a SAW filter for color TV IF circuits. He is currently engaged in the development of microwave integrated circuits, especially stabilized GaAs FET oscillators using dielectric resonators.



Yoshikazu Murakami was born in Aichi, Japan, in 1950. He received the B.S. degree in physics from Nagoya University, Nagoya, Japan, in 1973.

In 1974, he joined the Sony Corporation Research Center, Yokohama, Japan. From 1973 to 1980, he worked on the development of microwave integrated circuits, especially stabilized GaAs FET oscillators using dielectric resonators. He is currently engaged in the development of microwave filters using epitaxial yttrium iron garnet.

A Low-Pass Prototype Network Allowing the Placing of Integrated Poles at Real Frequencies

DAVID S. G. CHAMBERS AND JOHN DAVID RHODES, FELLOW, IEEE

***Abstract*—This paper details a procedure by which a number of attenuation poles can be placed at differing frequencies, giving an asymmetric or symmetrical response, the only restriction being that the network must be physically symmetrical. If a number of poles are placed on one side of the passband, this technique can be used to greatly increase the selectivity of a filter on this side, while maintaining an equiripple passband response.**

There are four possible arrangements for these filters. They can have even or odd degree with an even or odd number of integrated poles. Only three of these are realizable in a symmetrical network and these possibilities are dealt with individually.

An example is given in the case of an odd-degree filter with an odd number of integrated poles placed at two frequencies on opposite sides of the passband.

Manuscript received March 8, 1982; revised August 25, 1982.

The authors are with Filtronic Components Ltd., Royal London Industrial Estate, Acorn Park, Charlestown, W. Yorkshire, England.

I. Introduction

A NUMBER of microwave filter specifications call for a very low passband loss allied to extreme selectivity on one side of the band, and still requiring rejection on the

0018-9480/83/0100-0040$01.00

other. The optimum solution to this problem is an asymmetric bandpass filter.

The technique described in this paper can be used to arbitrarily place poles across a frequency band to ensure that a practical filter possesses more rejection across that band than a conventional device.

The low-pass prototype integrated pole filter can be realized in a number of microwave media [1], and both waveguide and combline devices of this type have been manufactured.

The modified approximation problem is first solved yielding a generalized transfer function. The pole locations are then calculated in order so that an alternating pole synthesis can be carried out, into a network containing "tri-sections" of impedance inverters to realize integrated real frequency poles. Finally, a nontrivial example is given of a seventh-degree filter with three integrated poles placed at two different frequencies. The input admittance can also be realized in an extracted pole circuit [2], or the poles can be a combination of extracted and integrated types.

For example, a seventh-degree device with three poles placed at two different frequencies could be realized with either three integrated poles or two extracted poles and a central integrated pole.

II. The Modified Approximation

The ω-plane poles of the ordinary Chebyshev function

$$T_n(\omega) = \cos\left[n\cos^{-1}(\omega)\right] \tag{1}$$

all lie at infinity, and hence it is an entire function, since there are no singularities in the complex plane. Hence, if m_i of these poles are removed and constrained to lie at ω_i, a modified characteristic function

$$C_n(\omega) = \cos\left[\left(n - \sum_1^L m_i\right)\cos^{-1}\omega + \sum_1^L m_i \cos^{-1}\left(\frac{F(\omega)}{\omega - \omega_i}\right)\right] \tag{2}$$

is produced where L is the number of frequencies at which poles are to be placed.

For an equiripple performance with the maximum number of turning points in the passband, each $\cos^{-1}$ term must be shifted by its maximum value (π) in the passband. Hence, as ω goes from -1 to $+1$, so $F(\omega)/(\omega - \omega_i)$ goes from -1 to $+1$, and the two conditions

$$\left|\frac{F(\omega)}{\omega - \omega_1}\right|_{\omega = \pm 1} = \pm 1 \tag{3}$$

can be applied to ensure that the argument of the cos function shifts by $n\pi$ in the passband. Therefore

$$F(1) = 1 - \omega_i$$
$$F(-1) = 1 + \omega_i \tag{4}$$

giving

$$F(\omega) = 1 - \omega\omega_i$$

and this solution is unique since the bilinear function $F(\omega)/(\omega - \omega_i)$ is completely defined by the two conditions (3) and the position of the pole ω_i. The new characteristic polynomial is

$$C_n(\omega) = \cos\left[\left(n - \sum_1^L m_i\right)\cos^{-1}\omega + \sum_1^L m_i \cos^{-1}\left(\frac{1 - \omega\omega_i}{\omega - \omega_i}\right)\right]. \tag{5}$$

Now

$$|S_{12}|^2 = \frac{Y_e - Y_o}{(1 + y_e)(1 + y_o)} = \frac{1}{1 + \epsilon^2 C_n^2(\omega)} \tag{6}$$

and from this the even- and odd-mode input admittances (Y_e) and (Y_o) can be obtained since the zeros of $(1 + Y_e)$ can be obtained by arranging the poles of the transfer function in order of decreasing imaginary part and selecting the alternate poles [3]. Thus

$$\text{zeros of } \left[(1 + Y_e)(1 + Y_o^*)\right] = \text{zeros of } \left[1 + j \in C_n(\omega)\right]$$
$$\text{zeros of } \left[(1 + Y_e^*)(1 + Y_o)\right] = \text{zeros of } \left[1 - j \in C_n(\omega)\right] \tag{7}$$

and by multiplying out the left-half plane zeros of $1 + j \in C_n(\omega)$ the numerator of $(1 + Y_e)$ can be found.

$$\text{NUM}(1 + Y_e) = \prod_1^{n/2} (p + \sigma_r + j\omega_r)$$

$$\text{where } \sigma_r, \omega_r \text{ are real and } \sigma_r > 0. \tag{8}$$

Now Y_e is a reactance function and can, therefore, be expressed as the ratio of an odd to an even polynomial, i.e.,

$$Y_e = N1/N2 \quad \text{and} \quad 1 + Y_e = \frac{N1 + N2}{N2}. \tag{9}$$

Y_e can now be calculated by forming the ratio of the even and odd parts of the right-hand side of (8), and Y_o can be found similarly using the complex conjugates of the remaining poles. Thus if

$$\prod_1^{n/2} (p + \sigma_r + j\omega_r)$$
$$= p^{n/2} + (a_1 + jb_1)p^{n/2-1} + (a_2 + jb_2)p^{n/2-2} + \cdots,$$
$$\sigma_r, \omega_r \text{ real; } \sigma_r > 0$$

then

$$Y_e = \frac{p^{n/2} + jb_1 p^{n/2-1} + a_2 p^{n/2} + \cdots}{a_1 p^{n/2-1} + jb_2 p^{n/2-2} + \cdots}. \tag{10}$$

III. General Synthesis Procedure

A. The Prototype Network

These filters can be realised using the prototype network shown in Fig. 1 [4]. Synthesis begins with the removal of a shunt capacitor followed by an admittance inverter, and this process is repeated until it is desired to remove an integrated pole. This is usually done towards the end of the synthesis, i.e., near the middle of the filter.

The real frequency poles can be removed using the

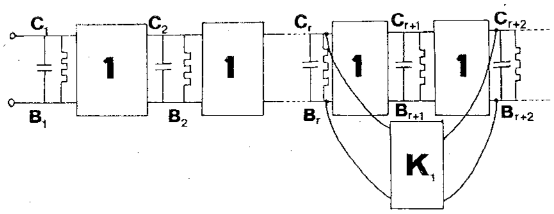


Fig. 1. Prototype network for integrated pole filters.

cross-coupled circuit in Fig. 2. First, however, an equivalent circuit suitable for direct synthesis must be derived.

The first stage in finding an equivalent circuit for Fig. 2 is to reflect the parallel resonant circuit in the first immittance inverter, Fig. 3(a). This gives a series resonant circuit where

$$B'_{r+1} = \frac{-1}{B_{r+1}} \tag{11}$$

followed by a $1:-1$ transformer in the main arm.

If the cross-coupling immittance inverter is now broken up into its Pi-equivalent circuit of frequency invariant reactances, the circuit of Fig. 3(b) results. Now, by partial fractioning the series impedance, the element values of Fig. 3(b) can be obtained from the circuit that is actually synthesized in Fig. 3(c) as

$$B'_{r+1} = \frac{-B'''^{2}_{r+1}}{B'''_{r+1} + B''_{r+1}}, \qquad K_1 = -B'''_{r+1}$$

$$C_{r+1} = \frac{B'''^{2}_{r+1}2}{C'_{r+1}}. \tag{12}$$

The $1:-1$ transformer scales through the network to the center line. Each integrated pole can be removed using one of these "tri-sections."

To begin the synthesis, a shunt capacitor is extracted

$$\frac{p}{a_1} + \frac{j\left(\frac{b_1}{a_1} - \frac{b_2}{a_1}\right)p^{n/2-1} + \cdots}{p^{n/2-1} + j\left(\frac{b_2}{b_1}\right)p^{n/2-2} + \cdots} \tag{13}$$

followed by an impedance inverter. This process is repeated until it is desired to remove an integrated pole. At this stage, a shunt frequency invariant reactance is removed in parallel with the shunt capacitor, so that the remaining even-mode impedance has a pole at ω_i

$$Y = \frac{ja_i p^i + a_{i-1}p^{i-1} + \cdots + a_0}{b_i p^i + jb_{i-1}p^{i-1} + \cdots b_0}$$

$$= jB + \frac{(ja'_{i-1}p^{i-1} + a'_{i-2}p^{i-2} + \cdots + ja'_0)(p - j\omega_i)}{b_i p^i + b_{i-1}p^{i-1} + \cdots jb_0} \tag{14}$$

where

$$jB = Y|_{p = j\omega_i}$$

and

$$a'_{r-1} = a_r + (-1)^{r+1}a'_r\omega_i + (-1)^r Bb_r.$$

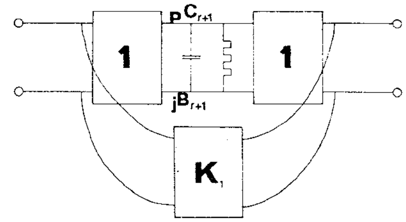


Fig. 2. An integrated pole tri-section.

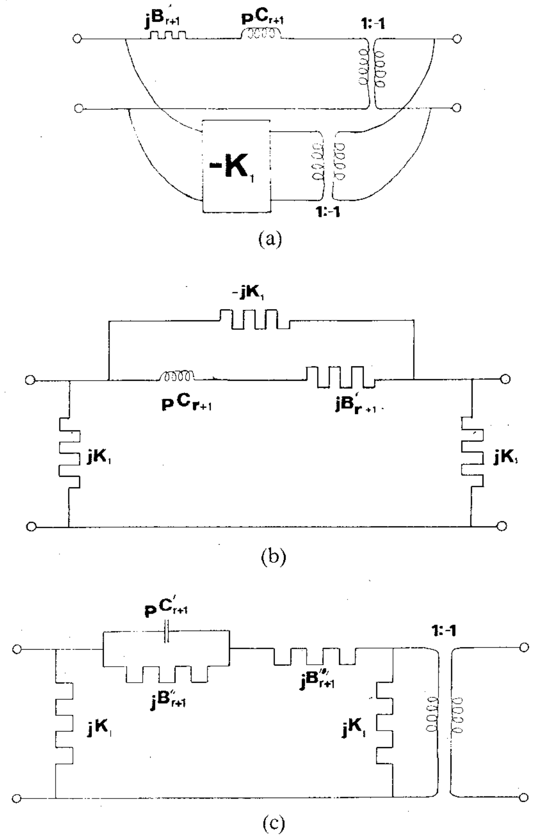


Fig. 3. Tri-section equivalent circuit derivation.

This pole at ω_i is then removed in the form of the resonant circuit of a capacitor and frequency invariant reactance

$$Z' = \frac{K}{p - j\omega_i} + \frac{b'_{i-1}p^{i-1} + jb_{i-2}p^{i-2} + \cdots + b'_0}{ja'_{i-1}p^{i-1} + a'_{i-2}p^{i-2} + \cdots + ja_0} \tag{15}$$

where

$$K = Z'(p - j\omega_i)|_{p = j\omega_i}$$

and

$$b'_{r-1} = b_r - a'_r K + (-1)^r \omega_i b'_r.$$

The remaining input admittance should be left with a pole at infinity in order that the next shunt capacitor can be removed, and this condition is achieved by removing a series frequency invariant reactance

$$ja'''_{i-1} + \frac{a'''_{i-2}p^{i-2} + a'''_{i-3}p^{i-3} + \cdots + ja'''_0}{p^{i-1} + jb''_{i-2}p^{i-2} + \cdots + b''_0}$$

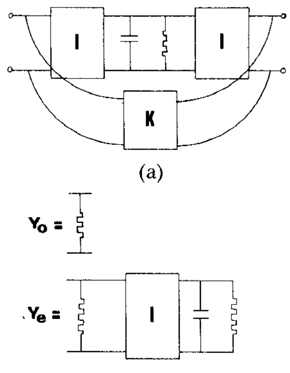


Fig. 4. Even- and odd-mode admittances of tri-section.

where

$$b_r'' = (-1)^r \cdot \frac{a_r'}{a_{r-1}'} \tag{16}$$

and

$$a_r''' = \frac{b_r'}{a_{r-1}'} + \frac{b_{r+1}' \cdot a_r'}{a_{r-1}' \cdot a_{r+1}'}.$$

The synthesis can proceed by removing a capacitor in parallel with a frequency invariant reactance followed by an impedance inverter. This can be repeated until the synthesis is completed or until another integrated pole needs to be removed.

To proceed with the synthesis, the three realizable network forms for these filters must be considered separately.

B. Even-Degree Filter with Even Number of Integrated Poles

In this case, both Y_e and Y_o are of degree $(n/2)$ and both the even- and odd-mode admittance provide the same circuit after synthesis. The only difference is in the value of the final frequency invariant reactances caused by the central admittance inverter. When synthesizing Y_o, this inverter has a short-circuit plane through its center, and when synthesizing Y_e, an open circuit, hence the value of this admittance inverter can be obtained by taking the difference between the final frequency invariant reactances in the two networks.

C. Odd-Degree Filter with Even Number of Integrated Poles

Here the odd-mode admittance Y_o has degree $(n-1)/2$ and Y_e has degree $(n+1)/2$. The m poles can be removed in pairs using the equivalent circuits derived in the previous section. In this case, however, the even-mode admittance will contain one more admittance inverter and capacitor than the odd-mode admittance; still the value of the central admittance inverter of the network can be obtained by taking the difference between the final frequency invariant reactances in each of the two circuits.

D. Odd-Degree Filter with Odd Number of Integrated Poles

Again, the odd-mode admittance has degree $(n-1)/2$ in this case, and the even-mode admittance has degree $(n+1)/2$. The first $(m-1)$ poles can be removed in pairs as in the previous case (with n odd and m even). In order to remove the final pole the even- and odd-mode equivalent circuits for Fig. 2 must be derived. As shown in Fig. 4, these circuits are then included in the relevant even- and odd-mode networks.

IV. Example of Synthesis Procedure

In order to demonstrate the synthesis procedure, consider a filter covering the band 7.25–7.75 GHz, which must provide 60-dB rejection over the transmit band 7.9–8.4 GHz, but due to local transmitter interference, it is required to maintain 50 dB at 7 GHz.

The low-pass prototype network required to meet this specification in waveguide has degree 7, and possesses two attenuation poles at $\omega_1 = 1.611790$ and a third pole at $\omega_2 = -1.965460$ (Fig. 5). These frequencies are normalized with respect to guide wavelength, and correspond to frequencies further away from the passband than the specification frequencies in order to ensure that the insertion loss characteristic meets the rejection specification on the monotonic portion of the insertion loss response. This minimizes the passband rolloff at the band-edge frequencies. This yields the characteristic polynomial

$$C_n(\omega) = \cos\left[4\cos^{-1}\omega + 2\cos^{-1}\left(\frac{1-\omega\omega_1}{\omega-\omega_1}\right) + \cos^{-1}\left(\frac{1-\omega\omega_2}{\omega-\omega_2}\right)\right]. \tag{17}$$

Using a computer program to find the zeros of $1 \pm j\in C_n(\omega)$ gives the left-half plane transfer function poles as

1) $-0.066497+j1.058027$
2) $-0.219670+j0.937238$
3) $-0.401932+j0.645414$
4) $-0.536840+j0.158690$
5) $-0.520479-j-0.411598$
6) $-0.348865-j-0.869225$
7) $-0.116467-j-1.096512$

and these are arranged in order of decreasing imaginary part.

In order to synthesize the even-mode input admittance, the alternate poles must be multiplied together and then the ratio of the even to the odd parts of the resulting polynomial can be formed. In this case, therefore, poles 1, 3, 5, and 7 should be multiplied together to form the even-mode input admittance, and poles 2, 4, and 6 give the odd-mode input admittance

$$Y_e = \frac{0.904670p^4 - j0.17671p^3 + 1.64695p^2 - 0.300709jp + 0.493941}{p^3 - j0.206339p^2 + 1.04847p - j0.201744}$$

$$= 0.904670p + \frac{j0.009959p^3 + 0.698431p^2 - 0.118197jp + 0.493941}{p^3 - j0.206339p^2 + 1.04847p - j0.201744}. \tag{18}$$

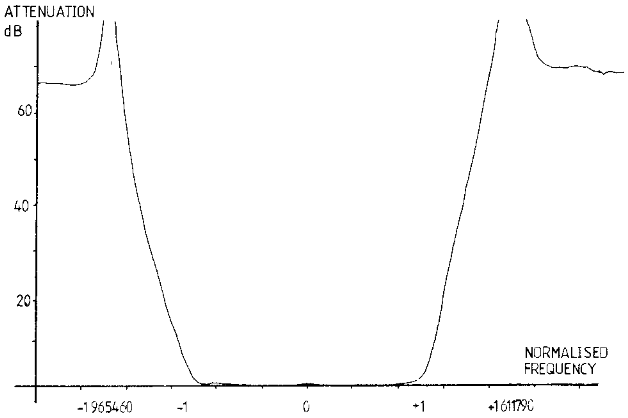


Fig. 5. Theoretical response of waveguide bandpass filter.

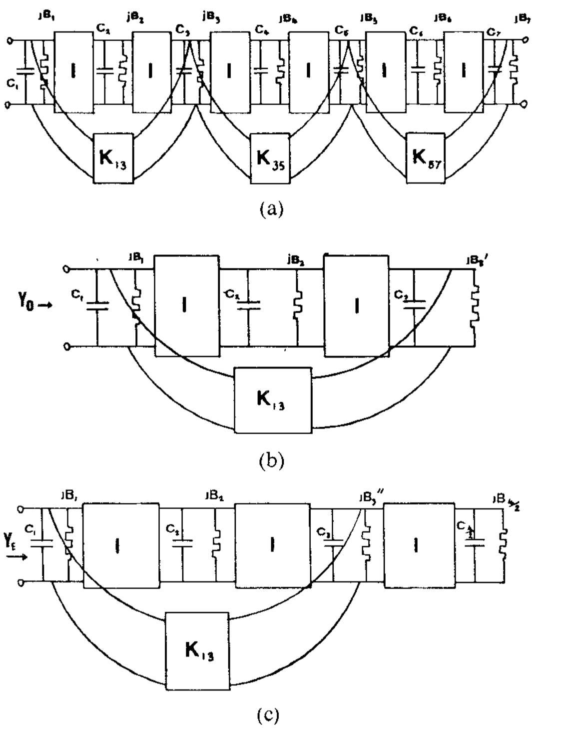


Fig. 6. Prototype network for seventh-degree filter with three integrated poles at two different frequencies.

This gives the value of C_1 in Fig. 6 as 0.904670. Next, a frequency-invariant reactance must be removed in order to leave the remaining admittance with a pole at ω_1, hence

$$Y_e' = jB + \frac{(p - j1.617790)(jAp^2 + Cp + jD)}{p^3 - j0.20633p^2 + 1.04847p - j0.201744}$$

where

$$jB = Ye'|_{p=j1.611790} = -j0.503133 \tag{19}$$

and A, C, and D can be found using (14). Hence

$$Z_e'' = \frac{p^3 - 5.0206339p^2 + 1.04847p - j0.201744}{(p - j1.611790)(j0.513092p^2 - 0.024750p + j0.369431)}$$

$$K = Z_e''(p - j1.611790)|_{p=j1.611790} \tag{20}$$

where E and F can be found from (15), giving $K = 2.155662$ and hence $C_2 = 0.463895$ and $B_2 = -0.747701$.

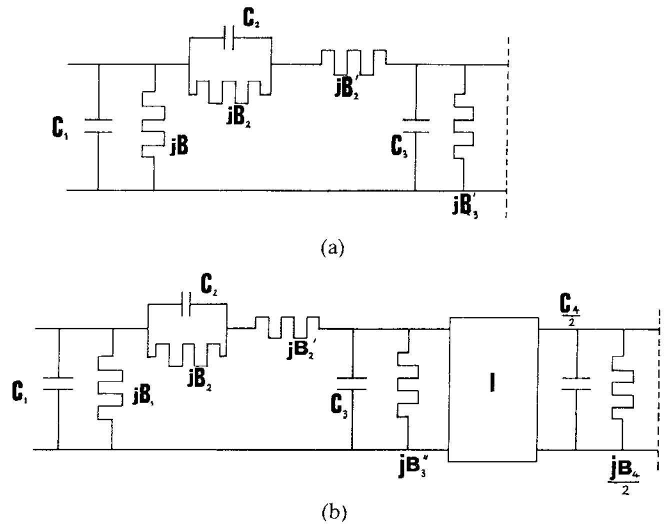


Fig. 7. Synthesized odd- and even-mode networks of the circuit shown in Fig. 6.

The synthesis continues by removing a series frequency-invariant reactance to leave the impedance with a zero at infinity

$$Z_e''' = \frac{-1.948968jp^2 + 0.583510p - 1.206910j}{p^2 + 0.048236jp + 0.720000}$$
$$= -1.948968j + \frac{0.489490p + 0.196340j}{p^2 + 0.048236jp + 0.720000}. \tag{21}$$

The remaining input admittance can now be synthesized using continued fraction expansion [5].

The odd-mode network can be synthesized from the complex conjugates of the remaining poles in a similar fashion. Using (11) and (12) to convert from the synthesized circuit results in the following values for the circuits of Fig. 6(b) and (c)

Y_o	Y_e
$C_1 = 0.904670$	$C_1 = 0.904670$
$C_2 = 1.762078$	$C_2 = 1.762095$
$C_3 = 2.042905$	$C_3 = 2.042942$
$B_1 = j0.009959$	$C_4 = 1.692380$
$B_2 = -j0.891147$	$B_1 = j0.009959$
$B_3' = j0.547548$	$B_2 = -j0.891150$
$K_{13} = 0.513096$	$B_3'' = -j0.207808$
	$B_4 = j0.67882$
	$K_{13} = 0.513092.$

To find K_{35} in the circuit of Fig. 6(a) use

$$K_{35} = \frac{B_3'' - B_3'}{2} = -0.377678$$
$$B_3 = \frac{B_3'' + B_3'}{2} = j0.169870. \tag{22}$$

The low-pass prototype values are now in the correct form for transformation into a suitable realization.

V. Conclusions

A design procedure allowing the construction of a transfer function containing arbitrarily placed attenuation poles has been presented, and it has been demonstrated how this transfer function can be synthesized into a low-pass prototype network containing integral cross coupling in the form of a "tri-section" of impedance inverters. Each tri-section realizes one integrated pole.

Comparisons made between conventional symmetrical techniques and filters built using the methods described here show that a significant reduction in degree can be obtained for a given selectivity by utilizing an asymmetric response.

An ordinary Chebyshev response can be recovered from the integrated pole transfer function by letting all the ω_i tend to infinity since

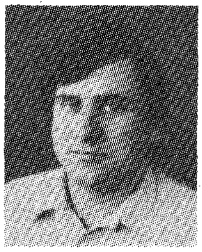
$$\lim_{\omega_i \to \infty} \frac{1-\omega\omega_i}{\omega-\omega_i} = \omega.$$

ACKNOWLEDGMENT

The authors wish to thank R. J. Cameron for his help in the numerical aspects of this work.

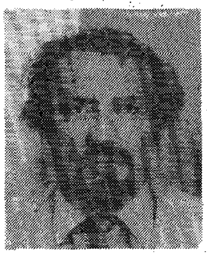
REFERENCES

[1] D. S. G. Chambers and J. D. Rhodes, "Asymmetric synthesis," in *Proc. 11th European Microwave Conf.*, Sept. 1981, pp. 105–110.
[2] R. J. Cameron and J. D. Rhodes, "General extracted pole synthesis technique, with applications to low loss TE_{011} mode filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-28, pp. 1018–1028, Sept. 1980.
[3] J. D. Rhodes and S. A. Alseyab, "The generalized Chebyshev low-pass prototype filter," *Int. J. Circuit Theory Appl.*, vol. 8, pp. 113–125, 1980.
[4] Richard M. Kurzrok, "General three-resonator filters in waveguide," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-14, pp. 46–47, Jan. 1966.
[5] J. D. Rhodes, *Theory of Electrical Filters*. New York: Wiley–Interscience, 1976.



David S. G. Chambers received the B.Sc. degree from the University of Leeds, Leeds, England.

Since 1980 he has been with Filtronic Components Ltd., Shipley, England, where he specializes in the research, development, and production of microwave filters and multiplexers.



John David Rhodes (M'67–F'80) was born in Doncaster, Yorkshire, England, on October 9, 1943. He received the B.Sc., Ph.D., and D.Sc. degrees in electrical engineering from the University of Leeds, Yorkshire, England, in 1964, 1966, and 1974, respectively.

He has worked as a Senior Research Engineer with Microwave Development Laboratories, Inc., Natick, MA, and has held positions of Lecturer, Reader, and Professor at Leeds University. In 1977 he formed Filtronic Components Ltd., Shipley, West Yorkshire, England, which has grown into a successful microwave company and he holds the position of Chairman and Managing and Technical Director, with a part-time position of Industrial Professor at Leeds University.

Dr. Rhodes has been the recipient of five international research awards including the Browder J. Thompson, Guillimen–Cauer Awards, and the Microwave Prize. He has also been a member of several professional committees.

Electronically Cold Microwave Artificial Resistors

ROBERT L. FORWARD, SENIOR MEMBER, IEEE, AND TERRY C. CISCO, MEMBER, IEEE

Abstract—**A large percentage of microwave field-effect transistors (FET's) are shown to act as a broad-band artificial resistor with a resistance of about 25 Ω when their drain is connected to their gate. The resistance appears between the gate–drain lead and the source lead. This resistance can be raised to 50 Ω with its reactive components eliminated over a reasonable bandwidth by using a matching transmission line of the proper impedance and a length near a quarter-wave at midband. An HFET-1000 constructed in this configuration showed an impedance of 18 ± 3 Ω over an octave bandwidth, and when transformed with a 30-Ω quarter-wave transmission line produced a resistance of 51 ± 1 Ω from 8 to 13 GHz. A noise analysis shows that, at some frequencies, some FET's in this configuration will produce artificial resistors with an effective noise temperature as low as 67 K. The addition of a transmission line in the gate lead and in the feedback line allows the effective temperature of any FET to be made substantially below room temperature. No cryogenic cooling is required. The addition of a half-wave to either transmission line will produce a low-noise negative resistor which can be used in a negative resistance amplifier circuit to produce an amplifier with the gain of a tunnel diode and the noise figure of a FET. These "electronically cold" artificial resistors could improve the overall noise figure of a microwave receiver if they replaced the first-stage amplifiers and the standard room-temperature loads in critical front-end circuits using loads in conjunction with circula-**

Manuscript received March 22, 1982; revised August 2, 1982.
R. L. Forward is with Hughes Research Laboratories, Malibu, CA 90265.
T. C. Cisco is with Hughes Missile Systems Division, Canoga Park, CA 91304.

0018-9480/83/0100-0045$01.00 ©1983 IEEE