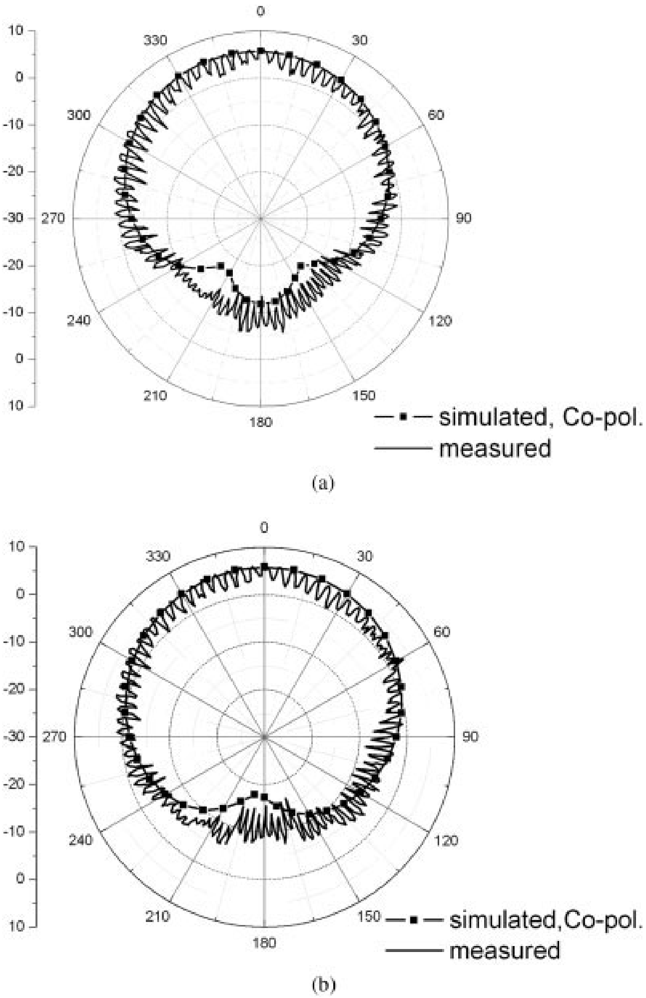

Figure 5 Simulated and measured radiation patterns at 4.21 GHz: (a) *x-z* plane and (b) *y-z* plane

unequal crossstub [4]. The difference between simulated and measured frequency range is probably due to that there is an air gap between the DRA and the substrate, which slightly lowers the equivalent permittivity of the DRA, thus lead to an increasing of frequency and expand the bandwidth [5]. The tolerance in the dielectric constant of DRA and dimensions would also contribute to the shift in the measured frequency.

Figure 4 shows the simulated and measured axial ratio against frequency and angle. The simulated bandwidth (AR $\leq$ 3 dB) of 53 MHz, covering the frequency range from 4.173 to 4.225 GHz is obtained with the minimum AR of 0.25 dB at 4.195 GHz, whereas the measured axial-ratio shows a little shift which covering the frequency range from 4.187 to 4.237 GHz with the minimum AR is 0.8 dB at the 4.21 GHz. Also, the axial ratio is found to be below 3 dB over a beam width of $\pm 82°$ from the boresight.

Figure 5 displays simulated copolarization radiation patterns and measured polarization radiation patterns of *x-z* and *y-z* planes at 4.21 GHz, showing a fine agreement. It is observed that broadside field patterns are obtained, as expected. The antenna gain was also studied. It is found that the measured antenna gain is 5.5 dBi around $f = 4.21$ GHz.

3. CONCLUSIONS

A novel DRA element fed by the Y-shaped microstrip line to realize the CP radiation with a wide bandwidth has been proposed. The circular polarization can be easily obtained by changing the length and width of the shorted stubs. The antenna is compact in structure, as it consists of only a cylindrical DRA with radius of 4.5 mm and a substrate of $30\,\mathrm{mm} \times 30\,\mathrm{mm}$. Its measured impedance bandwidth is about 6.8%, which is about 6 times wider than that for the microstrip feed without stubs and 1.5 times wider than the one using the microstrip with the unequal crossstub. The measured minimum axial ratio is 0.8 dB, and the beam width is 164°. It is suitable for many movable terminals of satellite communications.

ACKNOWLEDGMENTS

This work was supported by the National High-Technology Research and Development (863) Program of China, under Grant No.2007AA12Z125 and the Shanghai Leading Academic Discipline Project under Grant No.S30108.

REFERENCES

1. C.-Y. Huang, J.-Y. Wu, and K.-L. Wong, Cross-slot-coupled microstrip antenna and dielectric resonator antenna for circular polarization, IEEE Trans Antennas Propag 27 (1999), 605–609.
2. N. Farahat, W. Yu, R. Mittra, and T. Koleck, Cross-shped dielectric resonator antenna analysis using conformal finite difference time domain method, Electron Lett 37 (2001),1105–1106.
3. K.W. Leung, W.C. Wong, and H.K. Ng, Circularly polarized slot-coupled dielectric resonator antenna with a parasitic patch, IEEE Antenna Wireless Propag Lett 1 (2002), 57–59.
4. L.-N. Zhang, S.-S. Zhong, and S.-Q. Xu, Dielectric resonator antenna element and subarray with unequal cross-stub, Microwave Opt Technol Lett 50 (2008), 2189–2191
5. G. Drossos, Z. Wu, and L.E. Davis, The air gap effect on a microstrip-coupled cylindrical dielectric resonator antenna, Microwave Opt Technol Lett 20 (1995), 36–40

LOG PERIODIC BANDSTOP FILTER

E. S. Sithole and J. A. G. Malherbe

Department of Electrical, Electronic and Computer Engineering, University of Pretoria, Pretoria, South Africa 0002; Corresponding author: jagm@up.ac.za

Received 6 January 2009

ABSTRACT: *Most conventional transmission line filters are commensurate structures, with all elements that contribute to the filter performance of equal length. In this article, a structure is described that makes use of a series of elements that vary in length and spacing according to a logarithmic function; this results in a series of transmission zeros at real frequencies. It is shown that excellent filtering performance can be obtained from a relatively simple construction.* © 2009 Wiley Periodicals, Inc. Microwave Opt Technol Lett 51: 2418–2420, 2009; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.24652

Key words: *filters; transmission zeros; log periodic*

1. INTRODUCTION

Conventionally, transmission line filters are based either on image parameter design methods [1] or make use of design through modern network theory [2]. These designs are commensurate line networks, i.e., all elements are one-quarter wavelength long at the centre frequency. The designs are based on Richards' transform [3]

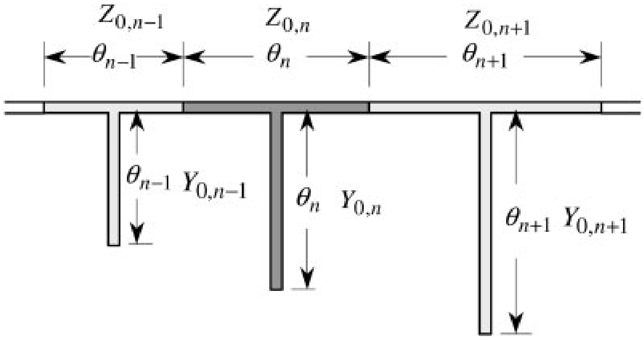

Figure 1 Configuration of the structure

that maps low-pass prototypes into bandstop structures, and high-pass prototypes to bandpass filters. An exception, but also based on Richards' transform, are the filters described in Ref. 4, where noncommensurate line are used. However, those structures are still based on Richards' transform, and also use the Kuroda and Kuroda-Levy transforms [2] to obtain physically realizable structures.

This article describes the properties of a structure that differs radically from those developed using conventional design procedures, in that it makes use of a series of stubs connected in shunt to a feed line, that all produce transmission zeros because of their individual resonant frequencies. The stub lengths and spacings are determined by a logarithmic function. While a complete design procedure is not presented, the useful properties of the structure are well illustrated.

2. LOG-PERIODIC STRUCTURES

It is shown in Ref. 5 that the reflection coefficient of the structure of Figure 1 can be approximated by the expression as follows:

$$\Gamma = \exp\left[-j\frac{\pi\log\theta_N}{\log\tau} + \chi\right] \tag{1}$$

where θ_N is proportional to the frequency, τ is the period, and χ is a constant. Evans [6] finds the response function "extremely difficult to control by choice of τ and the impedance levels, as these affect cutoffs in a complementary manner." Indeed, a filter response is shown in Ref. 6 that does have sharp cutoff, but the stopband ripple level varies by about 20 dB.

The ratio between successive line (and stub) lengths is the period,

$$\tau = \frac{\theta_n}{\theta_{n+1}} \tag{2}$$

where the individual connecting line impedances are $Z_{0,n}$ and stub admittances $Y_{0,n}$.

In this form, the structure has $3N$ variables, viz the line lengths, stub admittances, and connecting line impedances. Unfortunately, information on appropriate values or relationships between the various variables does not exist, and substantial simplifications have to be made to illustrate the usefulness of the structure.

3. PROTOTYPE

In the structure to be studied, a number of substantial simplifications are made. The impedance values of the connecting lines are all chosen to be the same, at 50 Ω. Similarly, the admittances of the stubs are all made equal, so that, apart from the periodicity and number of stubs, the only design choice is the impedance level.

The choice of τ, N, and Y_0 determines the ripple levels and the bandwidth; by judicious choice of these variables, acceptable and useful filter responses can be obtained, even though this be by repeated analysis rather than by design. Analysis of a substantial number of prototypes has shown that, for a relative bandwidth of the order of 100%, the periodicity, $0.9 \leq \tau \leq 0.95$. The number of stubs, N, is chosen odd, and the centre stub made to resonate at the centre frequency of the filter. The various line lengths are calculated from (2), and the stub admittance Y_0 is chosen. Numerical analysis of the prototype structure is trivial, and by varying the three parameters, an acceptable response is rapidly found. In general, increasing the number of stubs increases the bandwidth; decreasing the stub admittance also increases the bandwidth, but has a strong effect on the stopband attenuation level.

With a value of $\tau = 0.92$, $N = 7$, and $Y_0 = 0.01$S (100 Ω), the response of S_{11} and S_{12} shown as broken lines in Figure 2, was obtained. The stopband ripples lie below −30 dB, and except for one peak at −33 dB, are at the same height.

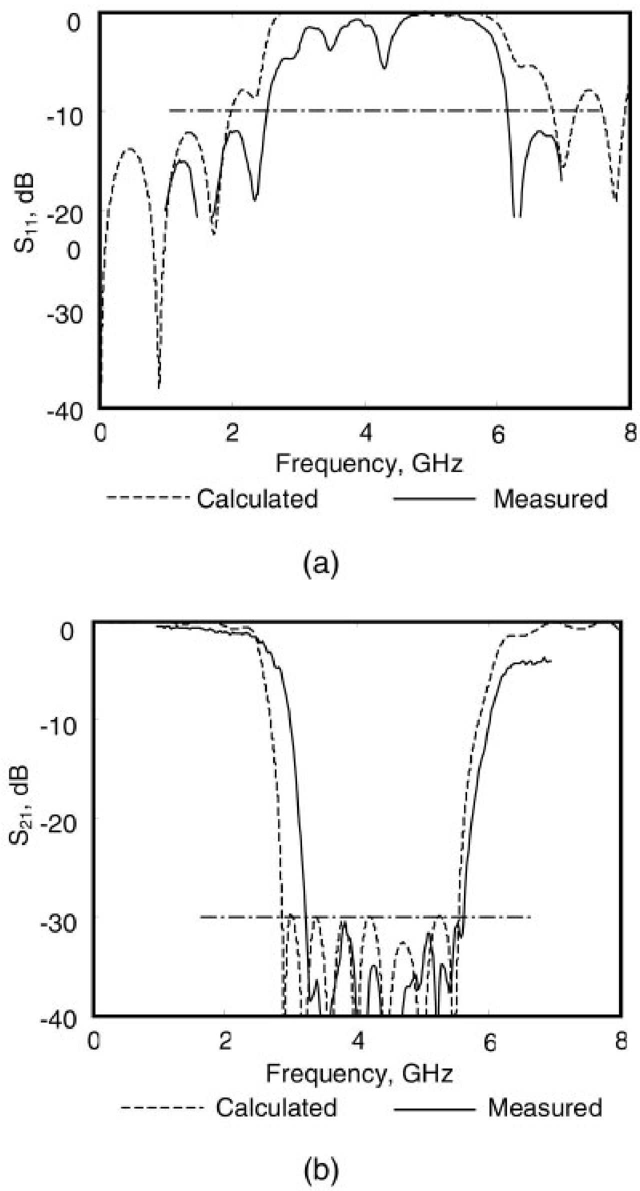

Figure 2 Responses of S_{11} and S_{21}

TABLE 1 Coupled Line Lengths

Stub No.	1	2	3	4	5	6	7
Length (mm)	14.30	12.89	11.52	10.45	9.33	8.31	7.56

4. PHYSICAL REALIZATION

The choice of the connecting lines to be the same at 50 Ω, and the stubs at 100 Ω, makes the filter very easily realizable by using coupled spurline sections [7]. All lines are 1.234 mm wide with gaps of 0.45 mm; the 50 Ω feed lines are 2.92 mm wide for a microstrip structure etched on FR4 substrate with a 1.6 mm dielectric thickness. The filter was designed for a centre frequency of 4 GHz, and the individual coupled line lengths are given in Table 1.

Figure 3 shows the steps in the development of the prototype filter, from stubs through spurlines. Also shown is the way in which the sections were folded so that the entire filter is realized as a section of line with etched slots. Because all the connecting lines are the same impedance, the spurlines can be constructed by using connecting line lengths from both the section itself, and the adjacent sections. Equal stub impedance furthermore makes it possible to have all the sections of the same cross-sectional dimension.

As an alternative, the filter could also have been realized simply as a 50 Ω line with stubs of 100 Ω; it is equally as realizable.

The measured responses for S_{11} and S_{12} for the prototype are also shown in Figure 2, where excellent agreement between the theoretical and measured results is observed. The measured results for reflection coefficient show discrepancies at singular frequencies, due to the fact that FR4 is not ideally suited to application to resonant structures at such high frequencies. The insertion loss due to the material is also quite apparent.

5. CONCLUSIONS

The usefulness of the log-periodic structure as a bandstop filter has been illustrated. While only a single filter at 100% bandwidth has been constructed and tested, calculated responses show that the range of applicability of the structure is extremely wide, with the order of the filter decreasing as the bandwidth decreases, and the stub impedance rising at the same time. For extremely wide responses (pseudo-lowpass), the number of stubs increases substantially, and both the pass and stop bands are degraded due to the limitations placed on the choice of both connecting and stub impedances.

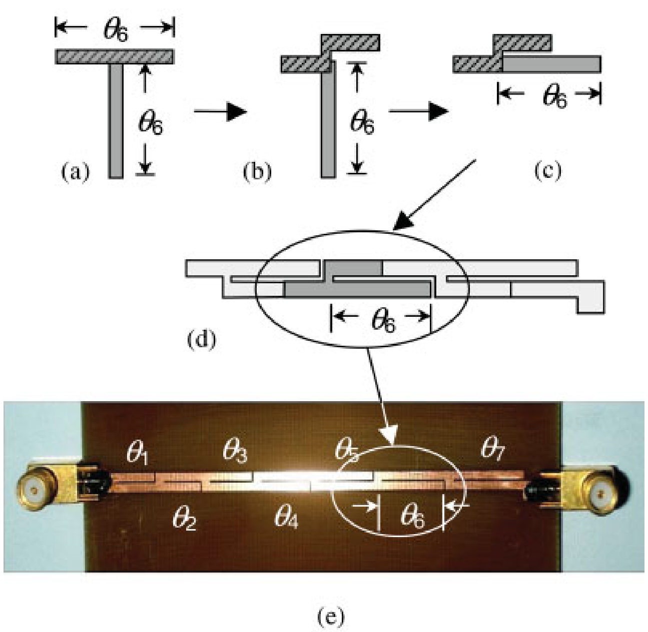

Figure 3 Steps in the development of the prototype filter. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

REFERENCES

1. G.L. Matthaei, L. Young, and E.M.T. Jones, Design of microwave filters, Impedance-matching networks and coupling structures, McGraw-Hill, New York, 1964.
2. J.A.G. Malherbe, Microwave transmission line filters, Artech House, Dedham, MA, 1979.
3. P.I. Richards, Resistor-transmission-line circuits, Proc IRE 36 (1948), 217–220.
4. J.A.G. Malherbe, TEM pseudo-elliptic function filters using noncommensurate lines, IEEE Trans Microwave Theory Tech 24 (1976), 242–248.
5. R.H. DuHamel and M.E. Armstrong, Log-periodic transmission line circuits. 1. One-port circuits, IEEE Trans Microwave Theory Tech 14 (1966), 264–274.
6. B.G. Evans, Performance of two types of log-periodic circuit, Electron Lett 5 (1969), 522–524.
7. B.M. Schiffman and G.L. Matthaei, Exact design of band-stop microwave filters, IEEE Trans Microwave Theory Tech 12 (1964), 6–25.

LOW POWER WIDE-LOCKING RANGE CMOS QUADRATURE INJECTION-LOCKED FREQUENCY DIVIDER

Sheng-Lyang Jang, Chia-Wei Chang, and Sheng-Ming Yang
Department of Electronic Engineering, National Taiwan University of Science and Technology, 43, Keelung Road, Section 4, Taipei, Taiwan 106, People's Republic of China; Corresponding author: m9502216@mail.ntust.edu.tw

Received 13 January 2009

ABSTRACT: *This letter presents a new low power and wide-locking range divide-by-2 injection-locked frequency divider (ILFD). The ILFD consists of a new 5.35 GHz quadrature voltage controlled oscillator (QVCO) and two NMOS switches, which are in parallel with the QVCO resonators for signal injection. The proposed CMOS ILFD has been implemented with the TSMC 0.18 μm CMOS technology and the core power consumption is 5.72 mW at the supply voltage of 0.8 V. The free-running frequency of the QILFD is tunable from 5.24 to 5.55 GHz. At the input power of 0 dBm, the divide-by-2 locking range is from 8.2 to 13.3 GHz as the tuning voltage is biased at 0.8 V. The phase noise of the locked output spectrum is lower than that of free running ILFD in the divide-by-2 mode. The phase deviation of quadrature output is about 1.28°.* © 2009 Wiley Periodicals, Inc.
Microwave Opt Technol Lett 51: 2420–2423, 2009; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.24640

Key words: *CMOS; locking range; divide-by-2; injection locking frequency divider; quadrature voltage controlled oscillator*

1. INTRODUCTION

Quadrature injection-locked frequency dividers (QILFDs) are free-running oscillators that take a sinusoidal input signal and generate four quadrature-phase signals at a frequency that is a fraction of the input signal. QILFDs can be used with a master voltage-