Microwave Self-Adaptive Filters and Set-On Oscillators for EW Systems.

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Abstract

This paper describes the development of a self-adaptive bandpass filter and a frequency set-on oscillator. The filter performs lowpass filtering on the instantaneous frequency of a signal, corresponding to a bandpass filter centred at the unknown carrier frequency of the signal. An experimental prototype working at 10GHz showed an increase in the received signal power by approximately 15dB while maintaining the noise floor level. When a feedback loop is added to a self-adaptive filter circuit, a frequency set-on oscillator is formed, which will sustain oscillations set up by an instruction pulse injected into the circuit. Seven modes of oscillation in the frequency range 9.85to10.05GHz were achieved during prototype testing, with an output power of approximately 13dBm. In this paper new design techniques for these filters and oscillators are presented, enabling significant reduction in system complexity.

Keywords: Adaptive filters, bandpass filters, delay filters, electronic countermeasures, microwave filters, radar receivers, frequency set-on oscillators, frequency memory loop, electronic countermeasures, microwave oscillators, radar receivers.

Motivation

Programmable adaptive filters and oscillators are useful in many areas of communications systems where accurate, stable selection and generation of tones is required. Applications of these filters are widespread in all types of communications systems. Applications include mobile telephony transceiver front ends, hearing aids [1], digital magnetic storage [2], broadband adaptive filtering in wired communications systems, echo cancellation, voiceband modems, and frequency memory loops in Electronic Warfare (EW) systems. This paper investigates adaptive filtering techniques, set-on oscillators and their application in EW systems. In all of these applications, systems that are robust, less expensive, easily implemented and incorporated into existing systems are favoured [3].

Introduction

An adaptive filter is a self optimizing system which is able to control its response to maximize the signal to noise ratio (SNR) in a communications channel [4]. Considerable work has been done on digital implementations of such devices, although they are limited to bandwidths of less than 100MHz [5]. Programmable or adaptive oscillators are useful in many different communication applications where accurate generation of RF tones is required. Electronic warfare systems need to determine carrier frequencies of incoming signals so that countermeasures may be deployed, or the source of these signals can be deceived by manipulation of the reflected signal. Fast locking oscillators are used to achieve this, the frequency set-on oscillator (FSO) described in this paper develops a device that may be deployed in Electronic Counter Measures (ECM) or in commercial communication systems.

Additionally, if these systems could be reduced in size, cost and power consumption they could be relocated from the transmitter to the receiver, with the inherent advantages that would bring [6]. EW systems work over multiple octave bandwidths; hence, they require an analogue solution.

An arbitrary connection of non-linear system components (i.e. mixers, frequency translators, etc) is termed a linear frequency network (LFN). Examples reported in [7], demonstrated a self-adaptive bandpass filter (SAF) and a FSO working over the range 400 to 600MHz. Advances in commercially available microwave components have meant that practical realizations of this concept can now be built at higher frequencies.

Some of the advantages of these techniques include; lower power consumption (which reduces heat dissipation issues), small size, greater reliability, lower cost, larger bandwidths, with lower complexity, increased stability and improved SNR. Monolithic Microwave Integrated Circuits (MMIC) are populated with both analogue and digital components describing a complete system, for these reasons detailed models of systems and subsystems are required. The goal of this project is a detailed investigation of the modelling and physical realisation of a SAF and FSO.

In this paper the work on the SAF and FSO has been significantly extended from that presented in [7], by adding more complexity to the system, it is possible to operate the filter in the 10GHz frequency range, while the majority of the components are specified for lower frequencies. In the next section, a review of the basic theory is presented. This is followed by a description of an experimental prototype system and measured results.

Theory

Certain non-linear devices produce linear transformations on the IF of a signal. Most importantly, linear frequency networks obey the principle of superposition in the Instantaneous Frequency (IF) domain [8, 9]. For a signal of the form $v(t) = a(t)\cos\left[\omega_0 t + \theta(t)\right]$, the IF is defined as

$$\omega_{IF}(t) = \omega_0 + \frac{d\theta(t)}{dt}.$$

If we consider the SAF circuit previously reported in [7], the transfer function G(p), can be defined by the two associated delays in the system, τ and τ_1;

for $z = e^{-p\tau}$ and $z_1 = e^{-p\tau_1}$,

$G(p) = H(z,z_1)$.

The transform of the input is simply $W_i(p)$, at the output of the frequency divider it is $W_i(p)/2$. When both the divided and the divided and delayed signals are present at the input to the mixer the output becomes $\frac{1}{2}(1+z)W_i(p)$. With the inclusion of the additional delay associated with the band defining filter, τ_1 (necessary for the prevention of anti-aliasing),

$H(z,z_1) = \frac{1}{2}(1+z)z_1$. The term $z_1 = e^{-p\tau_1}$ may be removed from this analysis (i.e. the inclusion of a delay in the circuit has no effect on the magnitude of the signals within the circuit). Giving the response of the filter as,

$$\left|H(z,z_1)\right|^2_{p=j\omega} = \cos^2\left(\frac{\omega\tau}{2}\right).$$

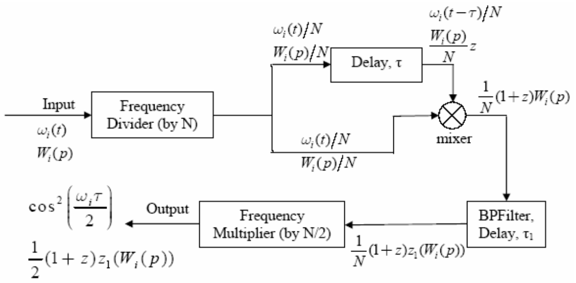

Figure 1: Schematic diagram of the SAF in the time and IF domains, divide by N approach.

Figure 1 extends this approach to using a divide by N block at the input: The input is still $W_i(p)$, but now the output from the divider is $W_i(p)/N$. The output from the mixer becomes,

$\frac{1}{N}(1+z)W_i(p)$. When N is an even integer, this system is practically realizable, because analogue multipliers can only produce integer harmonics of the input fundamental frequency. Including the z_1 term associated with the filter,

$H(z,z_1)=\frac{1}{N}(1+z)z_1$. We know that for the system H(z) to be stable $|z|\le 1$ and $|z_1|^2=1$ giving,

$H(z,z_1)\big|_{p=j\omega}=e^{\frac{-j\omega\tau}{2}}\left(\frac{2}{N}\cos\frac{\omega\tau}{2}\right)$. Now $\left|e^{\frac{-j\omega\tau}{2}}\right|^2=1$ and multiplying by N/2 gives the quasi-lowpass characteristic describing the system, as before

$$\left|H(z,z_1)\right|^2_{p=j\omega}=\cos^2\left(\frac{\omega\tau}{2}\right). \quad (1)$$

Equation 1 defines a filter with a $\cos^2$ profile centred on the D.C. component of the IF of the signal in the frequency domain.

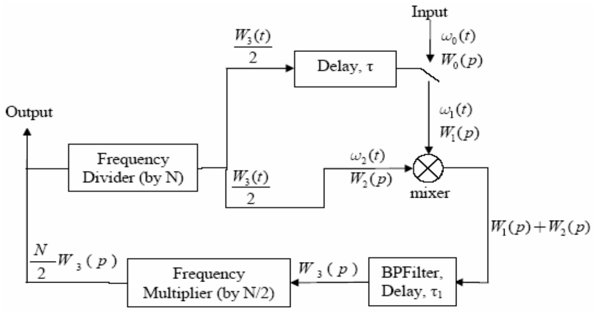

Figure 2: Schematic diagram of FSO in the IF domain, employing the divide by N approach.

A FSO is created when a feedback loop is added to the SAF circuit described in [7, 10]. The short term transient behaviour of the system is described thus [10],

$$\omega_2(t)=\omega_0+\psi_0\sum_{i=1}^{\infty}\partial(t-i\tau).$$

After many recirculations of the signal within the system described in Figure 2 the signal becomes;

$$\omega(t)=\omega_c+\sum_i\frac{b_i}{2}\cos\phi_i\sin\left(\frac{i4\pi t}{2\tau_1+\tau}+\psi_i^n\right),$$

where

$$\phi_i=\frac{i\pi t}{2\tau_1+\tau}.$$

What this means is that as the number of recirculations increases

$\omega(t)\rightarrow\omega_c$, hence, stable oscillations will occur at the mode closest to the injected instruction pulse frequency.

Figure 2 shows the way in which this new approach may be used in the design of the FSO to produce the same results. The ability to divide the IF down to baseband frequencies, employing the divide by N/multiply by N/2 blocks, where all the components become simpler presents itself using this approach. The main section of this system may then be carried out using DSP techniques at baseband frequencies, removing the inherent restrictions of sampling high frequency signals.

Prototype Testing and Results

Preliminary models and results first presented in [11] are extended in this paper to include the divide by N approach and the FSO. What follows is an outline of procedures, experimental prototype results and possible extensions to that achieved to date. The SAF will be presented first followed by the FSO.

The Adaptive Filter

A prototype of the system was built using commercially available microwave components as building blocks, see Figure 4. The frequency translators were from Hittite corp; amplifier and switch, Minicircuits; mixer, Watkins Johnson; couplers, minicircuits but mounted in house; lowpass filter, Telonic (f_{cutoff} = 2.7GHz); the delay line was 'X' Meters of semi-rigid co-axial cable (where 'X' defines the delay time). The switch was used to control the signal pulse period to the input of the device, note that there are two frequency dividers (divide by 2 and divide by 4) each side of the switch. This gives the divide by N approach with N=8. Couplers were placed in strategic positions in the circuit so that signals in the circuit could be monitored as the prototype was debugged.

Figures 5, 6 & 7 show the output spectra of the SAF, the noise floor of the SAF signal shows the characteristic profile of this type of system. Note that the period of the ripples in the noise floor was used to confirm the operation of the system. The periodicity of the ripples is defined by the delay line length being used, frequency ripple $f_r = \frac{1}{\tau}$ Hz, $\tau \gg \tau_1$ so τ_1 has little effect on the ripple spacing, Figure 7 confirms this.

For the measurements (Figures 5, 6 & 7) the input signal consisted of a synthesized source at the centre frequency (f_0 @ 0dBm) fed directly into the frequency divider. Figures 5 & 6 show the effect of using different delay times; where the periodicity of the ripples in the noise floor may be observed to differ.

Figure 5 shows a period of approximately 10MHz, while the longer delay used in Figure 6 shows a shorter ripple period spacing of approximately 6MHz. Figure 7 shows the response of the system at X band (10GHz) where the components are all within their specified working frequency range, indicating a gain in the input signal of approximately 15dB, the 10MHz periodic ripple is very clear here.

To determine the 'transfer characteristics' of the system a set of tests was conducted using two tones at the input. Two synthesized sources (f_0 @ 0dBm & f_2 @ -10dBm) were applied through a power combiner and fed into the divider. Three values of f_0 frequency (corresponding to the three pairs of curves in Figure 8) were chosen 9, 10 and 11GHz, at each of these frequencies measurements were taken over a range of f_2, $f_0 - 50MHz \leq f_2 \leq f_0 + 50MHz$.

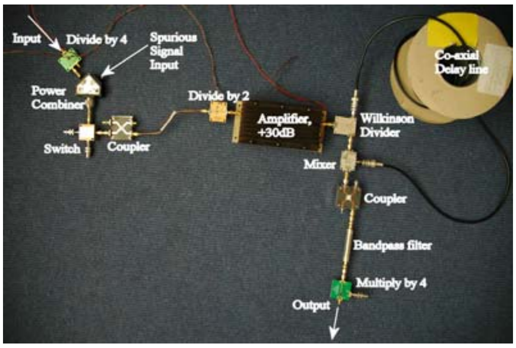

Figure 3: Photograph of the SAF experimental prototype.

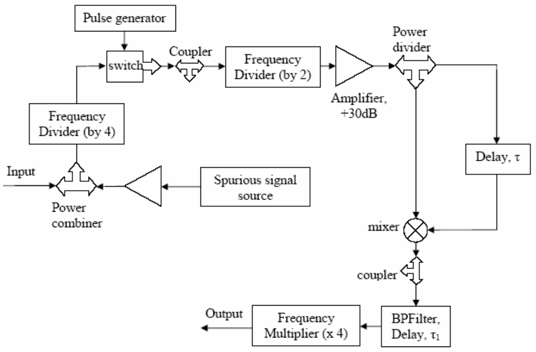

Figure 4: Schematic diagram of the SAF prototype.

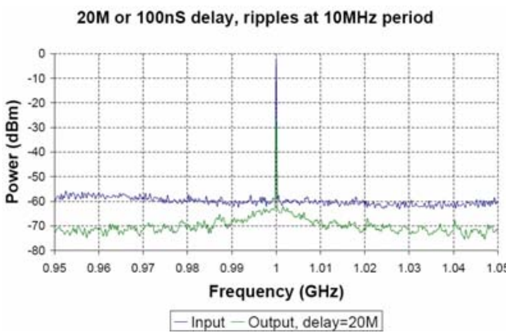

Figure 5: SAF output spectrum using a delay time of 100nS, tested at f_0 = 1GHz.

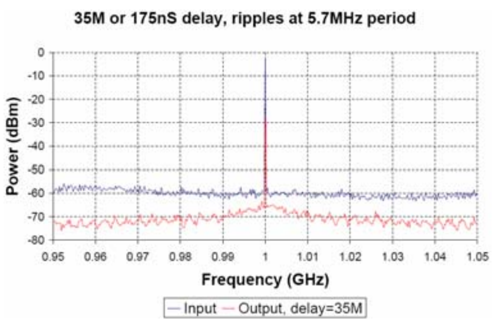

Figure 6: SAF output spectrum using a delay time of 175nS, tested at f_0 = 1GHz.

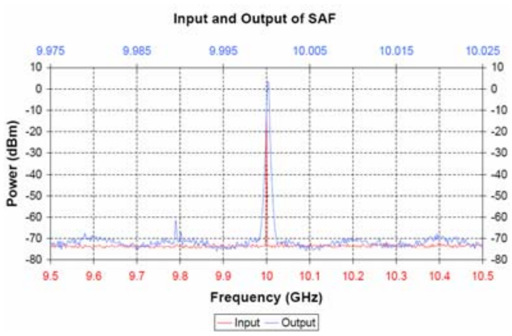

Figure 7: SAF input and output spectra, tested at f_0=10GHz.

The power gain at the frequency of f_2 (power at f_2 at the output in dB's - power at f_2 at the input in dB's) as a function of frequency difference from f_0 is shown in Figure 8. The graph shows a gain profile that resembles a $\cos^2$ form, indicating that the system attenuates spurious input signals. The interfering tone (f_2) power was significantly reduced in the f_0 = 9GHz test was due to the frequency multiplier (at the output) gain rolling off at this frequency. This was also evident in the f_0 = 11GHz test, but less profound. This set of measurements show how the bandwidth, gain profile flatness and linearity of each component in the system controls the overall performance of the system.

The Oscillator

Figure 10 shows the experimental prototype schematic while Figure 9 shows a picture of the FSO prototype. 10meters of semi-rigid co-axial cable was used to give the delay (50nS) which determines the modal spacing of the system, in this case,

$$f_{\text{modal}} = \frac{1}{\tau} = 20\,\text{MHz}.$$

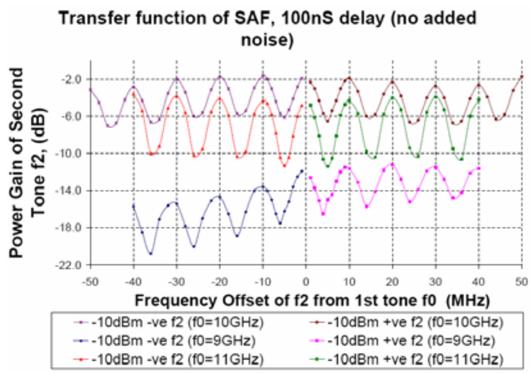

Figure 8: SAF two tone test results, spectra of the second tone gain.

Therefore, each oscillation mode is separated by 20MHz. The Hewlett Packard synthesized source (hp 8340A) was used to supply the instruction pulse; generated using a pulse generator and the pulse modulation input to the synthesizer. In this way the instruction pulse length could be varied. Again two frequency dividers were used (divide by 2 and divide by 4), this gives the divide by N approach with N=8. The 10GHz bandpass filter was custom designed and built in house.

The measurements presented in Figure 11 and 12 were achieved applying an instruction pulse, f_0 at -2.5dBm to port 1 of the coupler. Reducing the pulse length until the circuit failed to lock onto the instruction pulse, this gave the minimum pulse width. It was found to be approximately the length of the delay line - corroborating that reported in [7]. By varying the instruction pulse frequency from 9.85GHz to 10.05GHz in 2MHz steps it was possible to test the capability of the circuit to lock on to the closest mode.

Figure 12 shows seven modes of oscillation over the 200MHz bandwidth, separated by approximately 20MHz as defined by the delay line. The output power of the oscillator signal can be seen to vary between Figures 11 and 12, this is due to a slightly different set-up, the output signal in Figure 11 was taken after the band pass filter not after the feedback loop amplifier as in the figure.

It is clear that several modes were more prevalent than others, Figure 11 shows these modes. The more prevalent modes were located at frequencies where the loop gain was greatest; flattening this gain response will minimize these issues. Having determined that each component had a flat gain response, it is clear that the cause of the gain ripple and the decreased SNR lies in the reflections set up between components in the system i.e. it is necessary to reduce the mismatches in the circuit.

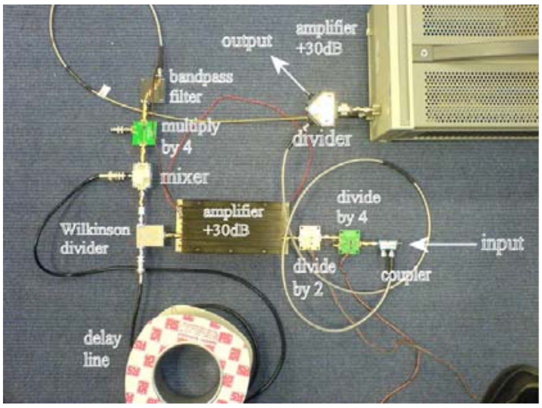

Figure 9: Photograph of the FSO prototype.

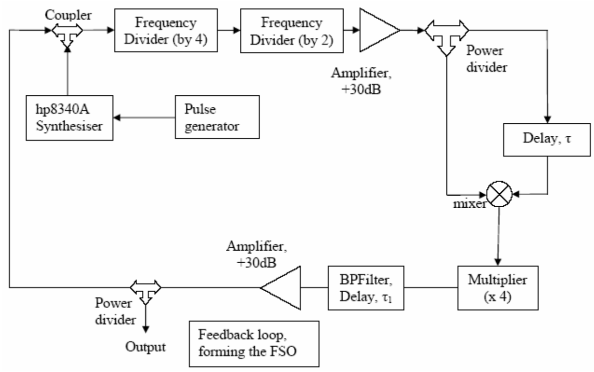

Figure 10: Schematic diagram of the FSO prototype.

The third peak from the left of Figure 12 (the thicker line) also indicates spurious reflections within the system, its separation from the other modes should be 20MHz yet it is only 10MHz from its adjacent peak. An additional reflection must exist in the delay line, creating a delay of 100nS and hence modal spacing of only 10MHz, again the need to reduce mismatches is clearly indicated.

Conclusions

This paper describes extensions to the design of self-adaptive filters and frequency set-on oscillators. By incorporating divide by N/multiply by N/2 circuits into the system it is possible to use much lower frequency components (and hence reduced cost).

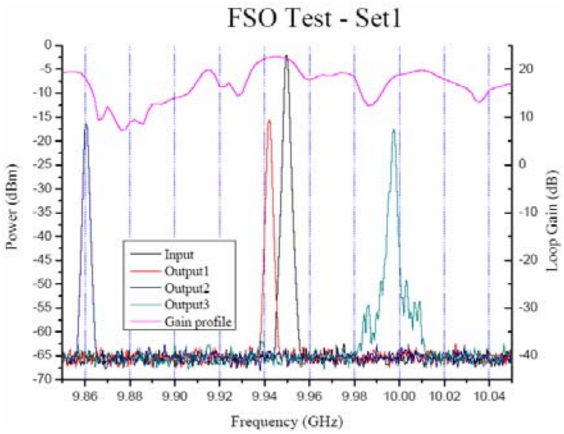

Figure 11: Graph highlighting the prevalent modes of oscillation in the FSO.

An experimental filter working at X band incorporating a divide by 8/multiply by 4 block has been developed. Measured results show that the filter places a bandpass characteristic around the carrier frequency, tested in the laboratory at 1 & 10 GHz (previous work was conducted from 400 to 600MHz) showed an improvement in the SNR of approximately 15dB.

Previously reported designs of FSO's have also been extended to include an arbitrary division factor. A prototype FSO working in the 10GHz range over a bandwidth of 200MHz has been developed using the divide-by-8/multiply-by-4 approach. Results show that seven distinct modes of oscillation with an output power of approximately 13dBm can be achieved over the specified band containing ten possible modes. The number of achievable modes within a band and the signal power can be improved by reducing the open loop gain ripple in the system.

Future Plans

In [11] the principals of linear frequency networks were explored, including a software model proving the concept. Now that working experimental prototypes have been developed, and extended to include the divide by N/multiply by N/2 approach, it would be useful to explore the possibility of employing DSP techniques in this system. The open loop gain in the FSO prevented the system from locking on to all possible modes in the band, by reducing the gain variation in the system this situation should improve making the system more suitable to its intended application.

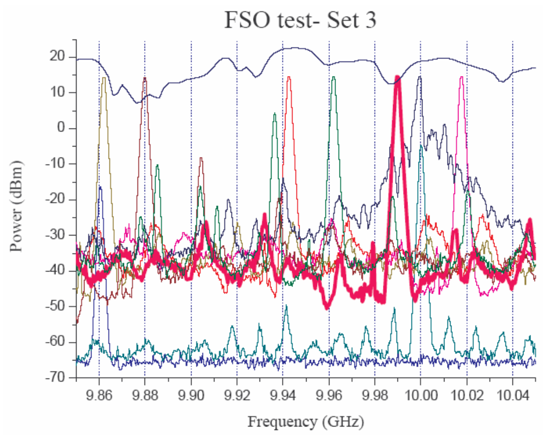

Figure 12: Graph showing the seven modal oscillations set up in the FSO.

Acknowledgements

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