Novel Detector Structure for Automatically Tuned Filters at Microwave Frequencies

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***Abstract* — In this paper a novel detector structure for an automatically tuned filter that is well suited for application at microwave frequencies is proposed. It can be used in a master-slave structure where the response of a master filter at a reference frequency is used to derive a control signal to steer a slave filter in the signal path. The control loop is analyzed theoretically and the result is compared to a measurement.**

***Index Terms* — Frequency control, tunable filters, microwave filters**

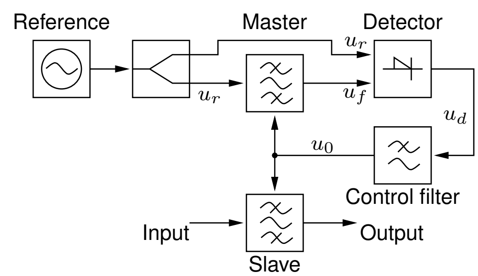

Fig. 1. Principle structure of the control loop.

I. Introduction

The characteristics of analog filters are vulnerable to fabrication tolerances and temperature drift. Automatic frequency control using the popular master-slave approach is the most feasible solution to control these characteristics [1]. It uses a master device that is embedded in a control loop to generate the tuning signal for the slave *Voltage Controlled Filter (VCF)* in the signal path (cp. Fig. 1). A good matching between master and slave is essential for a precise operation.

For microwave filters, this problem has been solved in [2], [3] by using a typical phase locked loop based on a voltage controlled oscillator matched to the slave filter.

Another control scheme that uses identical filters for master and slave was proposed in [4] and analyzed in detail in [5]. It utilizes a test signal to measure either the magnitude response (*Magnitude Controlled Filter, MCF*) or the phase response (*Phase Controlled Filter, PCF*) of the filter at a reference frequency. The analysis showed that the relative simple magnitude sensitive detector is not well suited because the system might hang up in an undesired operating point while a phase sensitive detector showed good performance at the price of higher complexity due to the necessary mixer.

In this paper a detector structure is presented and analyzed that shows a good performance without the need for a mixer.

II. Theoretical analysis

A. Analysis method

The structure of an automatically tuned filter following the master–slave principle is shown in Fig. 1. It was analyzed in detail in [5]. Here, only the most important results will be given briefly.

The sinusoidal reference signal $u_r(t) = \hat{u}_r \cos(\omega_r t)$ with angular frequency ω_r is applied to the master filter as well as to a detector. The behavior of the master filter with center angular frequency ω_0 is described by the transfer function $H(p;\omega_0)$. A second-order band pass transfer function will be used throughout the analysis:

$$H(p;\omega_0) = \frac{H_0 \dfrac{p}{\omega_0 Q_0}}{1 + \dfrac{p}{\omega_0 Q_0} + \dfrac{p^2}{\omega_0^2}} \qquad (1)$$

with H_0 being the gain at center frequency, $Q_0 = {}^{\omega_0}/_{\Delta\omega_0}$ and $\Delta\omega_0$ (in ${}^{\text{rad}}/_{\text{s}}$) the angular frequency bandwidth at $-3\,\text{dB}$. The relation between the center angular frequency and the tuning voltage is assumed to be linear for the moment. It is given by $\omega_0(t) = \omega_{0,0} + k_f u_0(t)$ with the VCF conversion factor k_f (in ${}^{\text{rad}}/_{\text{s V}}$) and the quiescent angular frequency $\omega_{0,0}$.

The output signal of the concatenation of the power splitter, the master filter and the detector is described by a function $u_d(t) = g(\omega_r, \omega_0, \hat{u}_r)$ which comprises the nonlinearity of the filter response as well as the behavior of the detector.

The *Control Filter (CF)* in the feedback path is typically a low pass filter together with an amplifier. In the analysis it is represented by its impulse response $h_{\mathrm{CF}}(t)$ or its Laplace transform, the transfer function $H_{\mathrm{CF}}(p)$. Thus

$$u_0(t) = \int_{-0}^{t+0} h_{\mathrm{CF}}(t-\tau)u_d(\tau)\mathrm{d}\tau + u_{0,0}(t) \tag{2}$$

with $u_{0,0}(t)$ considering settling effects due to initial conditions in the control filter.

The control system adjusts the center angular frequency ω_0 of the filter to the reference angular frequency ω_r. Its behavior is described by the integral equation:

$$\begin{aligned}\omega_0(t) = k_f \int_{-0}^{t+0} h_{\mathrm{CF}}(t-\tau) g\left(\omega_r(\tau), \omega_0(\tau), \hat{u}_r(\tau)\right) \mathrm{d}\tau \\ + \omega_{0,0} + k_f u_{0,0}(t)\,.\end{aligned} \tag{3}$$

Provided that the angular frequency and the amplitude of the reference signal are constant, the center angular frequency of the VCF will settle at a constant value. In terms of the system equation (3) this is equivalent to a steady-state solution. These stationary solutions are called *Operating Points (OP)*. The conditional equation for the OP might be derived from (3) for $(\omega_r,\, \omega_0,\, \hat{u}_r) = \text{const.}$:

$$\omega_0 = \omega_{0,0} + k_f \left(\lim_{t\to\infty} u_{0,0}(t) + H_{\mathrm{CF}}(0) g(\omega_r, \omega_0, \hat{u}_r)\right). \tag{4}$$

$H_{\mathrm{CF}}(0)$ is the DC-gain of the control filter. Usually, the limit of the control-filter transient-term tends to zero. For the stability analysis, the nonlinear function $g()$ of the system is expanded into a linear Taylor series at the OP:

$$g(\omega_r, \omega_0, \hat{u}_r) \approx u_{d,0} + K_d\omega_0 + K_r\Delta\omega_r \tag{5}$$

with

$$K_d = \left.\frac{\partial g}{\partial \omega_0}\right|_{\mathrm{OP}}, \qquad K_r = \left.\frac{\partial g}{\partial \omega_r}\right|_{\mathrm{OP}}, \tag{6}$$

$$\Delta\omega_r = \omega_r - \omega_{r,\mathrm{OP}}\,, \tag{7}$$

$$u_{d,0} = g(\omega_r, \omega_0, \hat{u}_r)|_{\mathrm{OP}} - K_d\omega_{0,\mathrm{OP}}\,. \tag{8}$$

Variations of the input amplitude $\hat{u}_r$ are not taken into account here.

The integral equation is linearized, Laplace transformed and rewritten as:

$$\omega_0(p) = \frac{\omega_{0,\mathrm{OP}}}{p} + \frac{K_r}{K_d} H_L(p)\Delta\omega_r(p) \tag{9}$$

with

$$H_L(p) = \frac{k_f K_d H_{\mathrm{CF}}(p)}{1 - k_f K_d H_{\mathrm{CF}}(p)} \tag{10}$$

to describe the system behavior for small disturbances from the OP. The well-known methods of system theory might be used to analyze the stability of each OP [6] that is determined by the stability of $H_L(p)$ for steady-state input signals.

B. Analysis of the detector

The novel detector structure is depicted in Fig. 2. It consists of two tunable master filters (1 and 2) that are detuned compared to the slave by voltages $u_{\varepsilon 1}$, $u_{\varepsilon 2}$. The power of the filtered reference signals is converted to proportional DC signals that are subtracted. For a linear tuning characteristic of the filters the detuning is proportional to the offset voltages:

$$\omega_{0,\nu} = \omega_{0,0} + k_f(u_0 + u_{\varepsilon,\nu}) = \omega_0 + \omega_{\varepsilon,\nu}\,, \tag{11}$$

$$\omega_{\varepsilon,\nu} = k_f\, u_{\varepsilon,\nu}\,, \qquad \nu \in \{1, 2\}\,. \tag{12}$$

Fig. 3 exemplifies the principle.

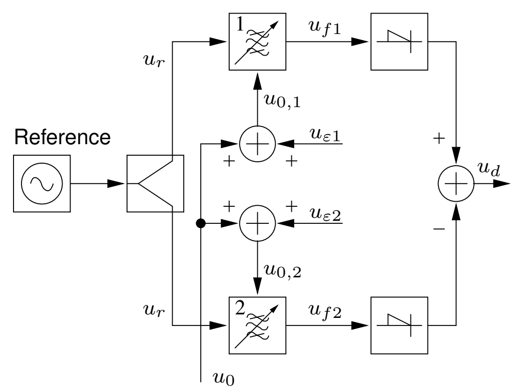

Fig. 2. Structure of the analyzed circuit. The slave filter is not shown. The nonlinear blocks represent power detectors.

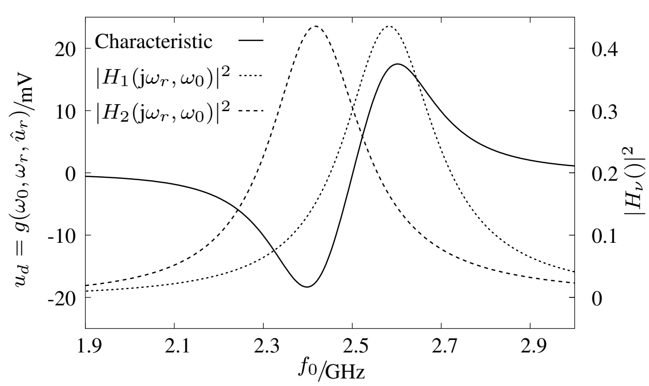

Fig. 3. Magnitude response of the detuned filters and resulting detector characteristic. The passband of the slave filter is around 2.5 GHz. Parameters: $f_0 = {}^{\omega_0}/_{2\pi}$, $f_r = {}^{\omega_r}/_{2\pi} = 2.5\,\mathrm{GHz}$, $H_0 = 0.66$, $Q_0 = 10$, $\omega_{0,0} = 2\pi \cdot 2.25\,\mathrm{GHz}$, $k_f = 1.3 \cdot 10^9\,{}^{\mathrm{rad}}/_{\mathrm{V\,s}}$, $k_p\hat{u}_r^2 = 62\,\mathrm{mV}$, $u_{\varepsilon,2} = -u_{\varepsilon,1} = 0.4\,\mathrm{V}$, $H_\nu(\mathrm{j}\omega_r, \omega_0) = H(\mathrm{j}\omega_r, \omega_0 + \omega_{\varepsilon,\nu})$.

The arrangement is similar to a balanced slope discriminator [7] that is used for the demodulation of frequency modulated signals with the difference that the input signal is at constant frequency here while the center frequencies of the filters are variable.

The characteristic of this detector is:

$$\begin{aligned} &g(\omega_0, \omega_r, \hat{u}_r) \\ &= k_p \hat{u}_r^2 \big(|H(\mathrm{j}\omega_r, \omega_0 + \omega_{\varepsilon,1})|^2 - |H(\mathrm{j}\omega_r, \omega_0 + \omega_{\varepsilon,2})|^2 \big) \\ &= \frac{k_p \hat{u}_r^2 H_0^2}{1 + Q_0^2 \left(\frac{\omega_0 + \omega_{\varepsilon 1}}{\omega_r} - \frac{\omega_r}{\omega_0 + \omega_{\varepsilon 1}} \right)^2} \\ &\quad - \frac{k_p \hat{u}_r^2 H_0^2}{1 + Q_0^2 \left(\frac{\omega_0 + \omega_{\varepsilon 2}}{\omega_r} - \frac{\omega_r}{\omega_0 + \omega_{\varepsilon 2}} \right)^2} . \end{aligned} \tag{13}$$

k_p is a proportionality constant of the power detectors.

If the control filter has integrating behavior:

$$H_{\mathrm{CF}}(p) = \frac{A}{p} \tag{14}$$

the control loop will settle in one of the zero crossings of the characteristic (eq. 4). For the practically relevant case of a small detuning ($\omega_r \gg |\omega_{\varepsilon,1,\varepsilon,2}|$) the only solution of interest is:

$$\omega_0 = \frac{1}{2} \left(-\omega_{\varepsilon 1} - \omega_{\varepsilon 2} + \sqrt{4\omega_r^2 + (\omega_{\varepsilon 1} - \omega_{\varepsilon 2})^2} \right) . \tag{15}$$

It is remarkable that if the frequency offset is chosen symmetrically around ω_0 (i.e. $\omega_{\varepsilon 1} = -\omega_{\varepsilon 2}$) the zero is not located at reference but there remains a frequency offset:

$$\omega_0 = \sqrt{\omega_r^2 + \omega_{\varepsilon 1}^2} . \tag{16}$$

This is because the amplitude response of the filter is not symmetrical with respect to the center frequency. However, the error is relative small: Assume the two master filters are detuned by the 3 dB bandwidth:

$$\omega_{\varepsilon 1} = -\omega_{\varepsilon 2} = \frac{1}{2} \omega_{3\,\mathrm{dB}} = \frac{\omega_0}{2Q_0} , \tag{17}$$

the center frequency is:

$$\omega_0 = \omega_r \frac{1}{\sqrt{1 + \frac{1}{4Q_0^2}}} \approx \frac{\omega_r}{1 - \frac{1}{8Q_0^2}} \qquad \text{for } Q_0 \gg 1 . \tag{18}$$

The relative error with respect to the 3 dB bandwidth is then approximately:

$$\frac{\omega_0 - \omega_r}{\omega_{3\,\mathrm{dB}}} \approx \frac{1}{8Q_0} . \tag{19}$$

E.g. for $Q_0 = 10$ the relative error is only 1.25% of the 3 dB bandwidth.

The stability of the linearized system is determined by the pole of the transfer function (eq. 10):

$$H_L(p) = \frac{k_f\, K_d\, A}{p - k_f\, K_d\, A} . \tag{20}$$

It is located in the left half of the complex plane if $k_f\, K_d\, A < 0$. The factor K_d describes the slope of the characteristic $g()$ in the operating point (eq. 6); its sign is independent from the reference frequency. Hence the sign of the product $k_f\, A$ can be chosen in the design to fulfill the stability criterion in the whole range of operation.

C. Nonlinear tuning characteristic

The proposed analysis might be extended to a system where the relation between the control voltage u_0 and the center frequency ω_0 of the filter follows a (monotonous) nonlinear function:

$$\omega_0 = v(u_0) . \tag{21}$$

This leads to a stretching of the curves along the f_0-axis in Fig. 3. Additionally the possible tuning range will be limited to a certain band with a realistic filter, but the principle shape of the characteristic will not change.

For the analysis the measured tuning characteristic of a filter was approximated by a square root function:

$$\omega_0 = v(u_0) = 2\pi \cdot 10^9 \frac{1}{\mathrm{s}} \left(c_0 + c_1 \sqrt{\frac{u_0}{\mathrm{V}}} \right) . \tag{22}$$

The amplitude response of the filters as well as the resulting detector characteristic are shown in Fig. 4. Around $f_0 = 2.5$ GHz the same detuning between the master filters is achieved as in the linear case. For higher frequencies it is reduced due to the flattening of the square root function. This leads to a reduction of the detector output signal compared to the linear case. Below 2.35 GHz the tuning range of one filter is exceeded.

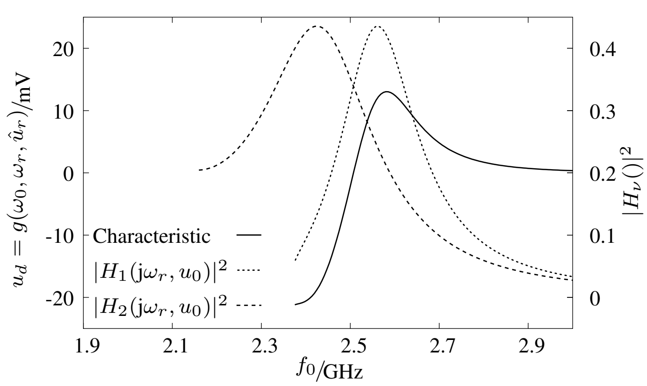

Fig. 4. Magnitude response of the detuned filters and resulting detector characteristic for a nonlinear tuning characteristic. The x-axis is converted to the corresponding center frequency of the slave. Parameters: $f_0 = \omega_0/2\pi$, $f_r = \omega_r/2\pi = 2.5$ GHz, $H_0 = 0.66$, $Q_0 = 10$, $c_0 = 2.1561$, $c_1 = 0.3392$, $k_p \hat{u}_r^2 = 62$ mV, $u_{\varepsilon,2} = -u_{\varepsilon,1} = 0.4$ V, $H_\nu(\mathrm{j}\omega_r, u_0) = H(\mathrm{j}\omega_r, v(u_0 + u_{\varepsilon,\nu}))$.

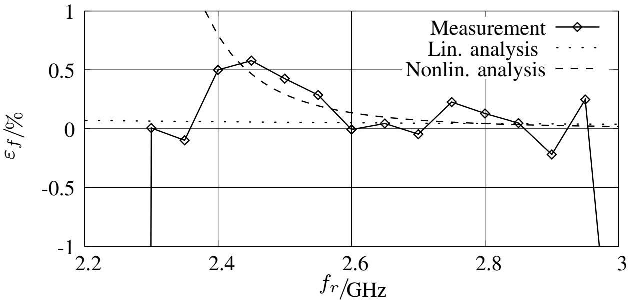

Fig. 5. Comparison of the relative frequency error between the measurement and the analytical results using a linear and a square root approximation for the tuning characteristic.

The analysis is carried out as described before. As the equation for the operating points can not be solved analytically due to the additional nonlinearity, it was solved using the computer algebra system *Maple 8* [8] for a given set of parameters. The result is shown in Fig. 5. This improved analysis predicts a larger frequency error than the analysis using a linearized characteristic but the error is still relative small compared to the 3 dB bandwidth. The stability analysis yields the same result as before: there is only one stable OP.

III. Measurements

A system was implemented to validate the analytical results. The filters were built up as first order parallel coupled filters with a varactor diode in the middle of the resonator using microstrip technology.

The filters are tunable from 2.2 GHz to 3 GHz. They have a quality factor of $Q_0 = 10$ and -3.6 dB damping in the passband at 2.5 GHz. The tuning characteristic of both filters showed a deviation of 20 MHz at maximum, which is less than 10% of the 3 dB bandwidth.

The parameter $k_p \hat{u}_r^2 = 62\,\mathrm{mV}$ that was used in the previous analysis was derived from the realized system at an input power level of -4 dBm with $f_r = 2.5\,\mathrm{GHz}$ and $u_{\varepsilon,1} = -u_{\varepsilon,2} = 0.4\,\mathrm{V}$.

The power detectors were built up using zero bias schottky diodes and a transmission line matching network. The integrator in the control filter was designed to have a unity gain crossover frequency of 15 MHz. It was implemented using operational amplifiers together with the summation network for the offset adjustment.

Fig. 5 shows the relative frequency error:

$$\varepsilon_f = \frac{f_0 - f_r}{f_r} \tag{23}$$

of the measurement together with the results of the theoretical analyzes. Obviously, the nonlinear model predicts the system's behavior much better than the analysis with a linearized tuning characteristic. In the range of roughly 2.3 GHz to 2.95 GHz, the relative frequency error of the controlled master filter is below 0.6% with respect to the reference frequency. This is approximately 6% of the 3 dB bandwidth of the filter and it can be regarded as sufficient for most applications.

An important parameter in the design of the detector is the detuning between the two master filters. A trade-off is necessary between different aspects: For a small detuning the output voltage is relative small which makes it more sensitive to noise. On the other hand both filters operate in the same region of the nonlinear tuning characteristic which leads to a smaller static frequency error. For a larger offset, the detector is less sensitive to tolerances between the master filters but the region of operation is reduced as one of the filters reaches the border of the tunable range earlier.

IV. Conclusion

In this paper an improved magnitude sensitive detector for an automatic frequency control system for tunable filters was analyzed. It showed a good performance without the need for a mixer. Measurements for a filter at microwave frequencies were presented that confirm the theoretical result.

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