A NEW DESIGN APPROACH FOR AN INJECTION-LOCKED OSCILLATOR WITH AN ENHANCED LOCKING RANGE

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ABSTRACT: *A new approach to the narrow locking range problems of an injection-locked oscillator (ILO) has been investigated using a feedback signal into the active device and a fictitious internal source generated by nonlinearities of the active device. Through this analysis, a new design methodology for an ILO with an enhanced locking range is proposed. Measured results verified the enhanced locking range of the ILO.* Microwave Opt Technol Lett 31: 325–327, 2001.

Key words: *injection-locked oscillator; locking range; frequency doubler; load pull*

INTRODUCTION

Nowadays, as the operating frequencies of microwave and RF transceivers are trending to shift to higher frequencies, local oscillators play an important role in communication systems. Phase-locked oscillators (PLOs) are commonly used in the systems, but they contain many components such as VCOs, phase detectors, and frequency dividers, resulting in a high cost. On the other hand, ILOs are simple, low cost, and millimeter-wave possible, but relatively do not have enough of a locking range [1, 2].

In this paper, we analyze an ILO in terms of feedback signal into the active device and a fictitious internal source generated by the nonlinear characteristics of the active device, and then propose a systematic design procedure to enhance the locking range of the ILO. Frequency doublers are designed for this procedure using the impedance substitute method (Fig. 1). Through this method, oscillators can be designed as amplifiers. That is, the feedback signals of the oscillators can be viewed as the input signals of the amplifiers with the same output powers. Therefore, the feedback signal levels into the transistors and the harmonic terminations maximizing the fictitious internal source can be determined easily in the design process. Compared with the conventional design procedure, this approach results in a wide locking range of the ILO.

DESIGN APPROACH FOR ILO

Figure 1 shows the impedance substitution method of the oscillator. jX_g, jX_s, and Z_d in Figure 1(a) represent the gate, source, and drain impedance, respectively, in the series feedback free-running oscillator topology. Also, $\Gamma_L(\omega_0)$ represents the impedance seen from between the source and drain toward the load. Figure 1(b) shows that we can substitute the amplifier topology for the oscillator topology if the operation

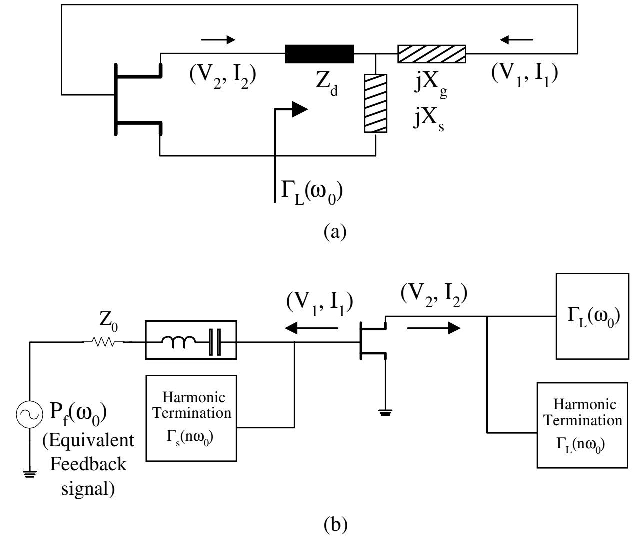

Figure 1 Impedance substitute method. (a) Series feedback oscillator topology. (b) Equivalent circuit of the oscillator in terms of amplifier topology

conditions of the active transistor (V_1, I_1), (V_2, I_2) are the same. In this case, $P_f(\omega_0)$ in Figure 1(b) represents the equivalent feedback signal which serves to drive the transistor. The advantage of this substitution method is that we can determine the feedback signal level (P_f) and the load condition (Γ_L) of the oscillator easily in the design process because we are generally accustomed to amplifier design.

Figure 2 shows a block diagram of the ILO. $Z_g(\omega)$, $Z_s(\omega)$, and $Z_d(\omega)$ in Figure 2(a) represent the gate, source, and drain impedance, respectively. Also, $P_f(\omega)$, $P_{inj}(\omega_{inj})$ represent the oscillation feedback signal into the gate and the external source, respectively. The three-port ILO can be

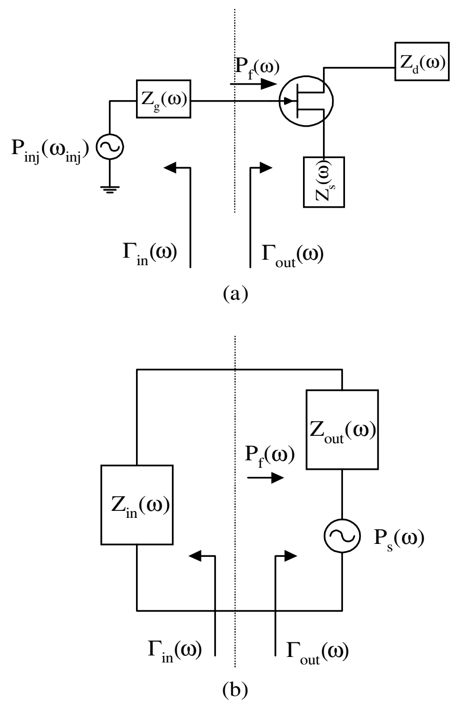

Figure 2 ILO block diagram. (a) Three-port ILO. (b) Two-port equivalent circuit with fictitious internal source

divided into two parts. One is $Z_{in}(\omega)$, and the other impedance is $Z_{out}(\omega)$ as in Figure 2(b), where $P_s(\omega)$ is a fictitious internal source generated in the transistor.

If the ILO is applied to the multiplier, the $P_s(\omega)$ is composed of the subharmonic component of the external source and the intermodulation components of two tones: the external signal and the feedback signal. In the case of a frequency doubler, the second subharmonic component of the external signal is the dominant component in $P_s(\omega)$. Therefore, $P_s(\omega)$ in the frequency doubler is independent of the feedback signal $P_f(\omega)$.

If no external source exists, that is, the steady-state free-running case,

$$P_f(\omega_0)\Gamma_{in}(\omega_0)\Gamma_{out}(\omega_0) = P_f(\omega_0) \tag{1}$$

where ω_0 is the free-running frequency and Γ_{in}, Γ_{out} are the reflection coefficients of Z_{in}, Z_{out}, respectively. If an external source whose frequency is $\omega_{inj} = (1/2)\omega_i$ is injected, the locking condition is satisfied as such:

$$\Gamma_{in}(\omega_i)[P_f(\omega_i)\Gamma_{out}(\omega_i) + P_s(\omega_i)] = P_f(\omega_i). \tag{2}$$

This equation is re-expressed by free-running frequency terms:

$$\begin{aligned}(\Gamma_{in}(\omega_0) &+ \Delta\Gamma_{in}) \\ &\times[(P_f(\omega_0) + \Delta P_f)(\Gamma_{out}(\omega_0) + \Delta\Gamma_{out}) + P_s(\omega_i)] \\ &= P_f(\omega_0) + \Delta P_f\end{aligned} \tag{3}$$

where ΔP_f, $\Delta\Gamma_{in}$, and $\Delta\Gamma_{out}$ are deviations of $P_f(\omega_0)$, $\Gamma_{in}(\omega_0)$, and $\Gamma_{out}(\omega_0)$, respectively, at the locking frequency. $\Gamma_{in}(\omega_0)\Gamma_{out}(\omega_0) = 1$, and if the load exists only in Z_d, $|\Gamma_{in}(\omega_0)| \approx 1 \gg |\Delta\Gamma_{in}(\omega_0)|$, $|\Gamma_{out}(\omega_0)| \approx 1 \gg |\Delta\Gamma_{out}(\omega_0)|$. Equation (3) is approximated as

$$(P_f(\omega_0) + \Delta P_f)\left(\Delta\Gamma_{out} + \frac{\Delta\Gamma_{in}}{\Gamma_{in}^2(\omega_0)}\right) + P_s(\omega_i) = 0. \tag{4}$$

In Eq. (4), the locking range of the ILO can be estimated because it is generally proportional to ΔP_f. The first term in the parentheses is related to the output power, the second to the Q of the circuit, and the last term means the nonlinear characteristics of the active device. They are similar to the well-known Adler's equation [3]. Therefore, in the case of constant Q, increasing the internal source and decreasing the feedback signal level can enhance the locking range. But the small feedback signal level in the design process lowers the output power.

DESIGN OF ILO

First, determination of the feedback signal level P_f is considered with respect to the output power and locking range because the output power and locking range are a tradeoff, as shown in Eq. (4). Then, the load pull of the free-running frequency is performed to search the load condition. Also, the load condition is related to the output power. If the feedback signal level and load condition are given, the port impedances of the free-running oscillator are determined uniquely as such [4]:

$$\begin{aligned}Z_d &= R_d + jX_d = z_2 + j\beta_b\frac{\mathrm{Re}\{z_1\}}{\mathrm{Im}\{\beta_b\}} \\ jX_s &= j\frac{\mathrm{Re}\{z_1\}}{\mathrm{Im}\{\beta_f\}} \\ jX_g &= j\,\mathrm{Im}\{z_2\} + j\,\mathrm{Re}\{z_1\}\frac{\mathrm{Re}\{\beta_f\}}{\mathrm{Im}\{\beta_f\}}\end{aligned} \tag{5}$$

where

$$z_1 = -\frac{V_1}{I_1}, z_2 = -\frac{V_2}{I_2}$$

$$\beta_b = -\left(1 + \frac{I_1}{I_2}\right), \beta_f = -\left(1 + \frac{I_2}{I_1}\right).$$

Second, harmonic terminations are needed to maximize the internal source. In order to search the harmonic load conditions easily, the external source $P_{inj}(\omega_{inj})$ can be added to Figure 1(b). Finally, RF input matching is considered for the external signal.

Two frequency doublers whose input frequencies are 5 GHz and whose output is 10 GHz were designed using the above procedure. One has a 5 dBm feedback signal without harmonic terminations, and the other has a 1 dBm feedback signal with harmonic terminations. But the two ILOs have the same Q, input VSWR, and output power. Table 1 shows the analytic results of port impedances of the ILO which has a 1 dBm feedback signal with harmonic terminations. And Figure 3 shows a schematic of the frequency doubler designed according to the values of Table 1. Bias circuits are omitted in Figure 3.

Figure 4 shows the experimental results of the locking ranges of the ILOs. As predicted, although Q and the output powers of the two ILOs are the same, the circuit with a small

TABLE 1 Port Impedance Values of the ILO

	Z_g (Ω)	Z_s (Ω)	Z_d (Ω)
@5 GHz	$13.4 + j66.4$	0	0
@10 GHz	$-j21.6$	$-j13.9$	$24.5 + j8.4$
@15 GHz	$362 + j277$	0	0

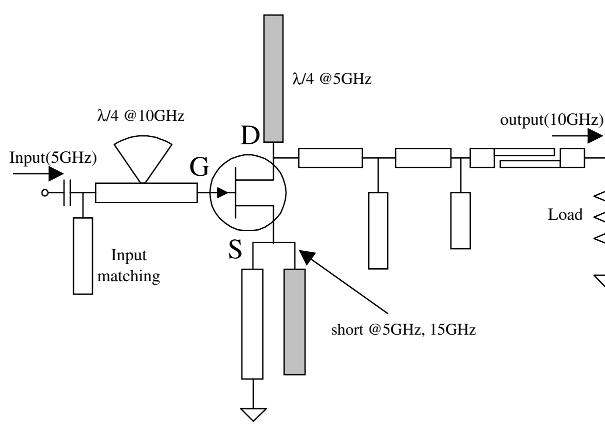

Figure 3 Schematic of the 10 GHz output frequency doubler

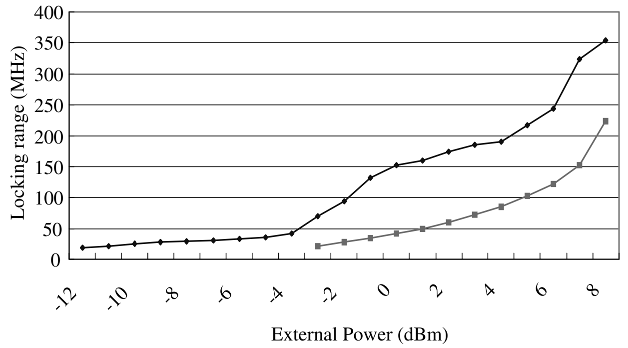

Figure 4 Locking ranges of ILOs (frequency doublers). –■– Frequency doubler which has 5 dBm feedback signal without harmonic terminations. –◆– Frequency doubler which has 1 dBm feedback signal with harmonic terminations

feedback signal and harmonic terminations had a much wider locking range.

CONCLUSION

A new design approach for an ILO with an enhanced locking range was proposed and applied to the frequency doubler. A small feedback signal level and harmonic terminations are key parameters for enhancing the locking range. These techniques show promise for lowering the complexity and cost of highly stable millimeter-wave source and phased-array antennas.

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BEF MADE OF COUPLED MICROSTRIP-LINE λ/4 RESONATORS

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ABSTRACT: *The present paper proposes a new design method of a miniatrized combline BEF. The degradation of the BEF characteristics due to close placement of the resonators is canceled with the aid of a capacitor between them. Simulation and experimental results based on the proposed design method are also presented.* © 2001 John Wiley & Sons, Inc. Microwave Opt Technol Lett 31: 327–329, 2001.

Key words: *band-elimination filter; combline; microstrip line; quarter-wavelength resonator; miniaturization*

1. INTRODUCTION

It was reported that the coupling effect between resonators leads to the degradation of attenuation performance for a combline-type band-elimination filter (BEF) with a strip-line structure [1]. A similar problem also arises in a BEF with a microstrip-line structure. Thus, in this paper, we propose a design procedure of the BEF to cancel the coupling effect. Since the coupled microstrip line has different effective dielectric constants for the even and odd modes, the design should be more complicated than for the strip-line structure.

2. DESIGN METHOD AND RESULTS

The following relations are satisfied for the coupled microstrip line shown in Figure 1(a):

$$\varepsilon_e > \varepsilon_o, Y_e < Y_o \tag{1}$$

where ε_e and ε_o are the effective dielectric constants for the even and odd mode, and Y_e and Y_o are the characteristic admittances for the even and odd mode, respectively [2]. Figure 1(c) shows the lumped-element equivalent circuit for the coupled microstrip-line $\lambda/4$ resonators shown in Figure 1(b). Here, C_e, L_e, C_o, L_o, L', and C' are

$$C_e = \pi Y_e/4\omega_e, L_e = 4/\pi\omega_e Y_e \tag{2}$$

$$C_o = \pi Y_o/4\omega_o, L_o = 4/\pi\omega_o Y_o \tag{3}$$

$$C' = (C_o - C_e)/2, L' = 2L_e L_o/(L_e - L_o) \tag{4}$$

where the resonant frequencies for the even and odd mode ω_e and ω_o are given as follows, respectively:

$$\omega_e = \frac{\pi}{2}\frac{c}{\sqrt{\varepsilon_e}\, l}, \omega_o = \frac{\pi}{2}\frac{c}{\sqrt{\varepsilon_o}\, l}. \tag{5}$$

Here, c is the velocity of light in vacuum and l is the length of the resonator. The relations in Eqs. (2)–(4) are obtained by the even- and odd-mode equivalence for Figure 1(b) and (c).

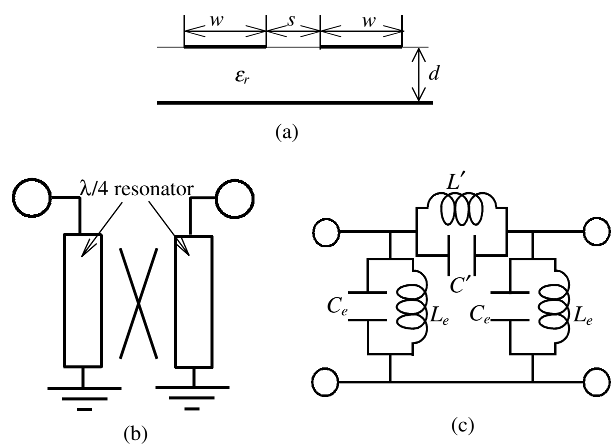

Figure 1 Coupled microstrip-line resonator. (a) Cross section. (b) $\lambda/4$ resonator. (c) Equivalent circuit of $\lambda/4$ resonator