COMPUTER-AIDED DESIGN OF PARAMETRIC AMPLIFIERS

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Abstract A computer-aided design approach is developed for the analysis and design of single-tuned parametric amplifiers. An 18 GHz single-tuned amplifier is designed and experimented for an illustration.

Introduction In this paper, a computer-aided design approach is developed for the analysis and design of single-tuned parametric amplifiers. The $\sqrt{GB}$ is used as a maximizing objective function for the design of the single-tuned parametric amplifier which consists of a cascade connection of distributed lines and a varactor diode. The design procedure has two parts: 1) the formulation of the $\sqrt{GB}$ in terms of circuit parameters; and 2) optimization using a modified pattern search maximization procedure[1].

Kinoshita and Maeda have reported an 18 GHz single-tuned parametric amplifier with a $\sqrt{GB}$ of 1450 MHz and a noise temperature of 245°K including a circulator loss at room temperature ambient[2]. The amplifier, using a pill-encapuslated varactor diode and a radial idler cavity, has been optimized experimentally. For an illustration, a single-tuned amplifier is designed and experimented at 18 GHz. The experimental amplifier alone exhibited a large $\sqrt{GB}$ of 2400 MHz and a low noise temperature of 180°K.

Design Formulation of Single-Tuned Parametric Amplifier A schematic of the parametric amplifier is shown in Fig. 1. $Z_s(f_s)$ and $Z_i(f_i)$ are the unpumped impedances of the signal and idler circuits, respectively. In order to operate as a parametric amplifier, the lossless coupling network should satisfy the following condition: 1) resonating the idler circuit at $f_{io}(X_i(f_{io}) = 0)$; 2) resonating the signal circuit at $f_{so}(X_s(f_{so}) = 0)$; 3) prohibiting the idler energy to propagate to the circulator $(R_i(f_{io}) = 0)$; and 4) holding the ratio $R_s(f_{so})$ to R_g at the value decided by a desired midband gain (for instance, $R_s(f_{so})/R_g = 1/(\tilde{Q}_1\tilde{Q}_2 - 1)$ under the large gain condition). The normalized $\sqrt{GB}$ of the single-tuned parametric amplifier at large gain is given by[3]

$$\frac{\sqrt{GB}}{f_{sc}} = \frac{2}{Q_s/\tilde{Q}_1\tilde{Q}_2 + f_{so}Q_i/f_{io}} \qquad (1)$$

where $\tilde{Q}_1$ and $\tilde{Q}_2$ are the dynamic quality factors of the varactor at f_{so} and f_{io}, respectively. Q_s and Q_i are the unpumped unloaded quality factors of the signal and idler circuits, and are given by

$$Q_s = \frac{1}{2R_s(f_{so})} \left. \frac{dX_s(f_s)}{df_s} \right|_{f_s=f_{so}} \qquad (2)$$

and

$$Q_i = \frac{1}{2R_s} \left. \frac{dX_i(f_i)}{df_i} \right|_{f_i=f_{io}} \qquad (3)$$

The objective here is to maximize $\sqrt{GB}/f_{so}$ under the conditions 1), 2), 3) and 4). Since the optimum idler frequency for the maximum $\sqrt{GB}/f_{so}$ is mainly decided by the self resonant frequency of a varactor, the noise temperature is sometimes deteriorated, compared with the minimum noise temperature obtainable with the varactor. For most practical parametric amplifiers, the noise temperature is one of the most important design factors and should be designed to be below the maximum allowable $(t_{e,\,m})$. Therefore, another design condition, i.e., 5) $t_e \leqq t_{e,\,m}$, must be added to the above conditions.

Typical Example of Coupling Network Figure 2 illustrates a typical configuration of the coupling network. The network consists of a pump line, an idler rejection filter, two transmission lines and a transformer. The lines are assumed lossless and uniform, and the transformer ideal. The equivalent circuit parameters associated with junctions between adjacent lines are disregarded in this example. Since the line 2 is the idler rejection filter and the idler circuit is composed of the line 1 and the varactor, the electrical length θ_2 is determined, independently of the characteristic impedance Z_2, from the condition 3) and the length θ_1 can be determined as a function of the impedance Z_1 from the condition 1). The line 3 is used for resonating the signal circuit. Once the parameters Z_1 and Z_2 are assumed, the electrical length θ_3 is obtained as a function of the impedance Z_3 from the condition 2). The condition 4) is satisfied by adjusting the ideal transformer ratio "n". Therefore, the independent variable parameters for this example are Z_1, Z_2, Z_3 and the idler frequency f_{io}, and $\sqrt{GB}/f_{so}$ can be expressed using these parameters.

Numerical and Experimental Results An 18 GHz single-tuned parametric amplifier was designed using above procedures. The equivalent circuit parameters of the varactor used (Sylvania D5147J, f_{c-6} = 600 GHz) are as follows:

$$C_{jo} = 0.14 \text{ pF} \quad C_p = 0.11 \text{ pF} \quad L_s = 0.15 \text{ nH}$$

The dynamic quality factors can be obtained experimentally by cold tests[4], and the typical measured value Q_1 at 18 GHz was about 4.5 including the circuit loss.

A modified pattern search maximization procedure was applied to the function (1). Several initial values of the parameters were then employed, because some discontinuites of $\sqrt{GB}/f_{so}$ resulted on the response hyper-surface. Figure 3 shows the calculated maximum $\sqrt{GB}/f_{so}$ and noise temperature versus the idler frequency f_{io}. The maximum $\sqrt{GB}$ is discontinuous when f_{io} is about 56 GHz. This discontinuity is based on the fact that the electrical length θ_1 determined from the condition 1) should be physically realizable, i.e., $\theta_1 \geqq 0$. It can be seen from Fig. 3 that the maximum $\sqrt{GB}$ can be obtained when the idler frequency is about 56 GHz.

A photograph and a measured bandpass characteristic of the experimental single-tuned parametric amplifier are shown in Fig. 4 and 5, respectively. The amplifier was mainly composed of coaxial lines except for a radial idler rejection filter and a pump waveguide. The idler frequency f_{io} was chosen at 55 GHz and the cut-off frequency of the pump waveguide at 60 GHz. As markers indicate the frequency differences of 50 MHz in Fig. 5, the 3 dB bandwidth B is 270 MHz at 19 dB gain and the $\sqrt{G}B$ is 2400 MHz. Noise temperature measurements indicated $180^{\circ}K$ for amplifier alone at room temperature ambient. On the other hand, the calculated $\sqrt{G}B$ and noise temperature are about 3000 MHz and $165^{\circ}K$, respectively. The differences between the measured and the calculated values are due to the fact that the calculation did not take into account the circuit parameters associated with the junctions between adjacent lines, nor the increase of the diode loss at the idler frequency. Moreover, the transformer was assumed ideal and its real frequency dependence property was neglected.

The forward pump power at 73 GHz required to drive the amplifier was about 30 mW after careful pump tuning. It is possible to use a solid-state pump source and simplify the amplifier.

Conclusion The computer-aided optimum design of the single-tuned parametric amplifier has been discussed. An 18 GHz single-tuned amplifier has been designed and experimented for an illustration. Although the design procedure has been developed for an 18 GHz parametric amplifier, the procedure can be applied to any parametric amplifiers.

The problems remaining to be solved are the consideration of parasitic elements associated with the line junctions, the measurement of the varactor equivalent circuit at millimeter-wave frequencies and the development of an efficient optimization algorithm for the problem which has some discontinuities on the response hyper-surface.

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References

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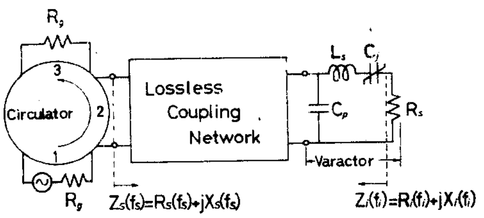

Fig. 1. Schematic of parametric amplifier.

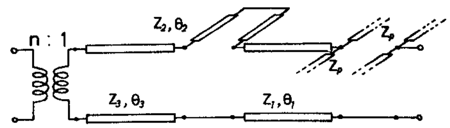

Fig. 2. Typical example of lossless coupling network.

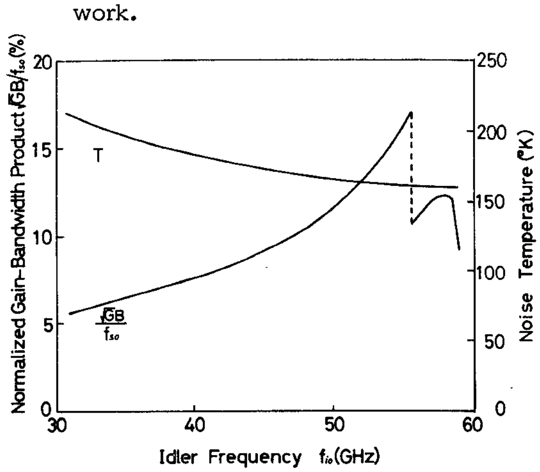

Fig. 3. Calculated normalized gained bandwidth product and noise temperature versus idler frequency.

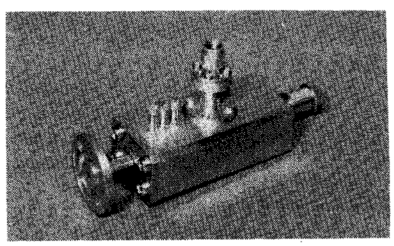

Fig. 4. Photograph of experimental single-tuned parametric amplifier.

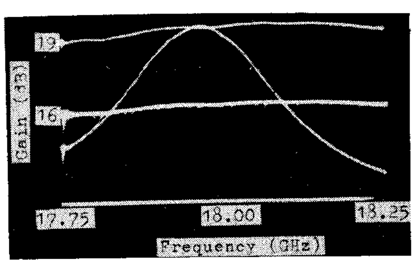

Fig. 5. Measured bandpass characteristics at 19 dB gain of experimental amplifier. Markers indicate the frequency differences of 50 MHz.