DESIGN AND EVALUATION OF A $g_m$-RC BANDPASS FILTER
USING A 42 GHz LINEAR OTA
INCORPORATING HETEROJUNCTION BIPOLAR TRANSISTORS

A Thesis Presented to
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CHAPTER 1
INTRODUCTION

There has been considerable interest in developing high frequency and microwave devices for integrated circuit realizations [1, 3-9]. Bandwidths of microwave signals used in signal processing applications are from 0.3GHz to 50GHz. Such high speed analog circuits demand high frequency response transistors. However, it is very difficult to have very high frequency response with Si bipolar transistor technology. Therefore, GaAs Heterojunction Bipolar Transistor (HBT) technology is being developed to meet the rapidly expanding market for faster devices. From a circuit viewpoint, although most traditional analog active circuits are implemented by conventional operational amplifiers, there are many restrictions in these amplifiers’ high frequency response. Therefore, Operational Transconductance Amplifiers (OTA) are being explored and developed in an attempt to do what the traditional Operational Amplifier can not normally do in high frequency applications. Thus, the superior high frequency performance of OTAs applying GaAs HBT technology is very useful in modern microwave and high speed communication engineering.

Computer simulation is a powerful supplement to traditional design techniques. In most design strategies, simulation can be an aid in the initial design development, during the breadboarding phase, and during the debugging and diagnostic phases. For some circuits, initial design theories must be tested before circuit design begins. With simulation, circuit blocks may be represented as functional elements and simulated in this form. Functional elements allow designers to test circuit theory without the time involved in developing transistor and component-level descriptions of each circuit function. For many circuits, breadboarding is impossible because of excessive circuit complexity, layout-
specific parasitic effects, or, as in the case of integrated circuits, both effects. For these types of circuits, simulation may be the only way to investigate the circuit behavior before building a working prototype. For most circuits, component value variation will have a direct effect on circuit performance and product yield. With simulation, designers can effectively predict the performance of a circuit as one or more circuit variables are changed. For all of these reasons, computer simulation is playing an increasingly important role in electronic circuit design.

In this thesis, we simulate the use of state-of-the-art GaAs Heterojunction Bipolar Transistors (HBTs) in a high-frequency Operational Transconductance Amplifier (OTA) and test the amplifier for a biquad $g_m$-RC bandpass active filter application. In addition, an accurate model of an OTA in the bandpass filter is developed. The model was able to predict the center frequency and estimate the $Q$ as well as predicting the effect of the OTA output load in the biquad $g_m$-RC bandpass active filter. All of the thesis work was performed by computer simulation tools. PSPICE[11, 29, 34], a powerful simulation program developed from the original SPICE program, was used to check our circuits and designs in this thesis.

The second chapter discusses the basic operating principles of HBT devices, and the advantages and disadvantage of HBT devices over silicon bipolar transistors. In the third chapter, basic characteristics of an OTA are brought out, a linear OTA is introduced, the quantitative analysis of the OTA is done, and the results of PSPICE simulations are given. Chapter four discusses the basic characteristics of second-order filter functions, introduces a biquad $g_m$-RC bandpass active filter utilizing three OTAs, and shows the results of PSPICE simulations with a simple OTA model and on the transistor level. In chapter five, the nonidealities of an OTA are explored and a proposed OTA model is presented. Simulations of the filter utilizing the model are shown. Chapter six presents a
modification of the filter, which has better simulation results. The limitation of the proposed OTA model also is discussed here. The last chapter consists of a summary of the thesis and suggestions for future work.
CHAPTER 2

HBT DEVICES

2.1 Introduction

Heterojunction bipolar transistors (HBTs) are the most mature of a new generation of III-V semiconductor transistors which rely on the use of heterojunctions for their operation. The heterojunctions in these devices are formed between semiconductors of different compositions and band gaps, e.g., GaAs/AlGaAs. This is in contrast to conventional Si- and GaAs-based field-effect (MOSFET) and bipolar devices which utilize junctions between like materials, e.g., n- and p-type Si in a bipolar transistor. These novel devices offer potential advantages in microwave, millimeter-wave, and high-speed digital integrated circuit (IC) applications over the homojunction devices presently in use [15-16, 19, 23]. This chapter discusses the basic operating principles of HBT devices, and the advantages and disadvantage of HBT devices over silicon bipolar transistors.

2.2 Basic Operating Principles of Heterojunction Bipolar Transistors

In conventional semiconductor devices only one type of semiconductor is used throughout, and control of current flow is achieved by creating a junction within the structure. This type of device is termed a homojunction. If more than one semiconductor material is used, causing a change in the energy bands within the structure, it is a heterojunction[22]. Fig. 2.1 shows energy band diagrams of a homojunction and a heterojunction. In the homojunction device, the barrier seen by the electrons injected from the emitter into the base and the barrier seen by holes injected from the base to the emitter are same. In the npn HBT, the barrier that electrons must overcome when they move from
the n side to the p side are quite different from the barrier for holes moving from p to n. The band offset $\Delta E_g = E_{gb} - E_{ge}$ at the heterointerface will favor injection of electrons, in an n (emitter) - p (base) heterojunction, from the emitter into the base while retarding injection of holes from the base into the emitter [15].

To demonstrate the operation of the HBT in the active region, the band structure of an n-AlGaAs (emitter) / p-GaAs (base) / n-GaAs (collector) HBT is shown in Fig. 2.3 (b). Here, the emitter-base junction is a heterojunction and the collector-base junction is a homojunction. The polarity connections are shown in Fig. 2.2. For the forward bias mode, electrons are injected from emitter to base, surmounting the potential barrier $qV_n$, while holes are injected from base to emitter, surmounting the potential barrier $qV_p$. As $qV_n$ and $qV_p$ satisfy the relation [33]

$$qV_p = qV_n + \Delta E_g$$

where $\Delta E_g$ is the band gap difference between AlGaAs and GaAs. The emitter injection efficiency $\gamma_e$ is defined by the ratio of the electron current (in the n-p-n transistors) which is due to the electron injection from the emitter to the total emitter current[24]. Thus,

$$\gamma_e = \frac{I_{Ee}}{I_{Ee} + I_{Ep}}$$

Obviously for high emitter efficiency, the back injected current from the base $I_{Ep}$ should be minimal. The ratio of the hole injection current $I_p$ to the electron injection current $I_n$ is calculated as follows [33]:
where \( v_{pe} \) and \( v_{nb} \) are the hole and electron average velocities at the emitter and base sides, respectively, and \( N_b \) and \( N_e \) are the majority carrier concentration of the base and emitter, respectively. Equation (2.3) implies that the value of \( I_p / I_n \) in an HBT could be decreased markedly when it is compared with the value of \( I_p / I_n \) in a homojunction bipolar transistor. This is because the term \( \exp(-\frac{\Delta E_g}{kT}) \) typically has a value of \( 10^{-5} \) in HBT devices. The small value of \( I_p / I_n \) leads to higher emitter injection efficiency which makes HBTs have advantages over their homojunction counterparts - Si bipolar transistors - as will be detailed below.
Fig. 2.1 Energy band diagram of (a) Homojunction and (b) Heterojunction.
Fig. 2.2 Definition of Terminal Voltages.
Fig. 2.3 Contrast of carrier injection at the emitter of (a) homojunction BJT and (b) a heterojunction bipolar transistor (HBT).
2.3 Advantages and Disadvantages of HBTs over Silicon Bipolar Transistors

In the homojunction BJT, once a material system is chosen (say Si, Ge, GaAs, etc.), the only flexibility one has in the device design is the doping levels and the device dimensions. In the HBT devices, the ability to tailor the energy-band structure adds flexibility to the design of new devices based on doping and material variations in the various layers. The changes in the energy band provide an additional means to control the flow and distribution of charge carriers throughout these devices [24].

In homojunction Si-bipolar transistors, it is necessary to use lightly doped material for the base region and heavily doped material for the emitter in order to maintain a high current gain $B$ because there is no band-gap difference. Unfortunately, the requirement of light base doping results in undesirably high base resistance. Thus a big advantage of an HBT is its much higher emitter efficiency. This allows the base to be more heavily doped than the emitter. The arrangement allows the designer to achieve a very thin base without increasing the base layer sheet resistance. This also allows one to have a relatively thick emitter-base depletion layer lying mostly inside the lightly doped emitter body. Therefore, there is a low emitter-base junction capacitance. These three features - high emitter injection efficiency, low base resistance, and low emitter-base capacitance - allow HBT devices to get high current gain at high frequencies [15-16, 19, 24]. Some other advantages of HBTs over silicon bipolar transistors are:

• Higher speed.
• Low pad parasitics to allow convenient integration of devices.
• Lower output conductance.

With the exception of higher device performance, HBTs behave very similarly to
their counterparts - silicon homojunction bipolar transistors. The circuit usage of GaAs HBTs therefore strongly resembles that of Si homojunction bipolar transistors[15]. Although Si bipolar transistors can be replaced by HBTs in many applications, cost considerations prohibit most such substitutions. The application of HBTs is limited to special areas where the performance of Si transistors is unacceptably poor and the cost of higher performance Si transistors exceeds that of HBTs. At present, this is usually the case for microwave frequencies above 1GHz[15-16, 19].
CHAPTER 3
THE OPERATIONAL TRANSCONDUCTANCE AMPLIFIER
(OTA)

3.1 Introduction

Although currently a large majority of active filters are built with voltage-controlled voltage sources, such as conventional operational amplifiers, it has become increasingly apparent that filters based on op amps are very restricted in their applications. Most importantly, the frequency-dependent gain of op amps tends to impose quite serious deviations on filter behavior and thereby precludes active filter applications significantly above the audio range[17]. Further, designers so far have generally not been successful in developing op amp-based active RC analog filters in monolithic form at high frequency and convenient voltage or current control schemes for externally adjusting the filter characteristics do not exist[18]. To combat these shortcomings, a great amount of effort is being directed towards designing active filters with operational transconductance amplifiers(OTA’s) in which the output current is controlled by an applied input voltage signal. Operational transconductance amplifiers generally have significantly higher bandwidths than op amps; in addition, they provide simpler circuitry for integration and easy methods for electronic tuning by changing a bias current. Finally, analog filters built with transconductances usually turn out to require fewer components than their op amp counterparts[10].

This chapter discusses basic characteristics of an OTA, introduces a specific tunable fully differential linear OTA circuit, presents the quantitative analysis of the specific OTA circuit, and shows the results of PSPICE simulations.
3.2 Basic Characteristics of an OTA

An ideal OTA is a voltage-controlled current source. The circuit symbol and the equivalent circuit of an ideal operational transconductance amplifier(OTA) whose input and output impedances are both infinite, are shown in Fig. 3.1. Its characteristics can be described by:

\[ I_o = g_m (V^+ - V^-) \]  \hspace{1cm} (3.1)

as illustrated in the diagram of Fig. 3.1 (b). In many designs, the transconductance \( g_m \) is adjusted by setting a control bias current, \( I_{ctn} \), such that \( g_m \) is proportional to \( I_{ctn} \). Many useful circuits have been developed based on the ideal transconductance model[8, 10]. However, when developing filters based on OTAs, the designer must take into consideration that real transconductances have finite input and output impedances. In many applications it is not the intrinsic and always present frequency dependence of \( g_m \) that imposes limits on transconductance operation, but rather the two time constants set by the input and output impedances in addition to any possible load effects[8].
Fig. 3.1 Circuit symbol (a) and ideal small-signal equivalent circuit (b) of a $g_m$ adjustable fully differential linear OTA circuit.
3.3 The Specific $g_m$ Adjustable Fully Differential Linear OTA Circuit

This thesis considers OTA circuits made out of GaAs heterojunction bipolar transistors. The PSPICE BJT model parameters of the HBT devices and the physical definitions of the BJT model parameters are shown in Table 3.1. All of the simulations in this thesis use the BJT model parameters for the HBT devices. The $i_C - v_{CE}$ characteristics of the BJT model of the HBT devices obtained from PSPICE simulations are shown in Fig. 3.2 and Fig. 3.3.

The schematic diagram of a specific tunable fully differential linear OTA circuit is presented in Fig. 3.4[8]. The linear OTA circuit consists of two main parts. One is the transconductance amplifier (Fig. 3.5). The other is the common-mode feedback (CMF) circuit (Fig. 3.6). These parts can be broken down further to constant-current sources, the differential pair, active load, biasing circuit, transconductance amplifier, and common-mode feedback circuit. We shall discuss the circuit structure and function of each part in the OTA circuit and present the quantitative circuit analysis.
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Table 3.1 The PSPICE BJT model parameters of the HBT devices and the physical definitions of the BJT model parameters.
Fig. 3.2  (a) Circuit for calculating the $i_C - v_{CE}$ characteristics of the BJT model of the npn HBT device with PSPICE. (b) The $i_C - v_{CE}$ characteristics of the BJT model of the npn HBT device as simulated by PSPICE.
Fig. 3.3  (a) Circuit for calculating the $i_C$ - $V_{CE}$ characteristics of the BJT model of the pnp HBT device with PSPICE. (b) The $i_C$ - $V_{CE}$ characteristics of the BJT model of the pnp HBT device as simulated by PSPICE.
The schematic diagram of a specific $g_m$ adjustable fully differential linear OTA circuit.
Fig. 3.5  Transconductance amplifier of the OTA.
Fig. 3.6  Common-mode feedback (CMF) circuit of the OTA.
3.3.1 Constant-current Sources

The ideal constant-current source is an electric circuit element that provides a current to a load that is independent of the voltage across the load. In electronic circuits, especially for integrated circuits, there are many applications for constant-current source[14,20]. Although it is never possible to have an ideal constant-current source in a real electronic circuit, there are ways to produce circuits that provides a very close approximation to the ideal constant-current source.

The current mirror, shown in its simplest form in Fig. 3.7, is the most basic building block in the design of IC current sources. The current mirror consists of two matched transistors Q₁ and Q₂ with their bases and emitters connected together, and which thus have the same \( V_{BE} \). In addition, Q₁ is connected as a diode by shorting its collector to its base. The current mirror is shown fed with a constant-current source \( I_{REF} \) and the output current is taken from the collector of Q₂. The circuit fed by the collector of Q₂ should ensure active-mode operation for Q₂ (by keeping its collector voltage higher than that of the base) at all times.

Assume that the BJTs have high \( \beta \) and thus that their base currents are negligibly small. The input current \( I_{REF} \) flows through the diode-connected transistor Q₁ and thus establishes a voltage across the base to emitter junction of Q₁ that corresponds to the value of \( I_{REF} \). This voltage in turn appears between the base and emitter of Q₂. Since Q₂ is identical to Q₁, the emitter current of Q₂ will be equal to \( I_{REF} \). It follows that as long as Q₂ is maintained in the active region, its collector current \( I_o \) will be approximately equal to \( I_{REF} \). This basic current mirror configuration serves as an active load of the differential amplifier in the OTA circuit considered here, which will be discussed in subsection 3.3.3.
Fig. 3.7  The basic BJT current mirror.
3.3.2 The BJT Differential Pair

The differential amplifier is a very important transistor amplifier stage configuration and is widely used in various types of analog ICs[12-14, 17-18, 27, 30]. For a general analysis of the BJT differential pair, we will consider the basic differential-amplifier circuit shown in Fig. 3.8.

If we denote the voltage at the common emitter by $v_E$, the exponential relationship applied to each of the two transistors may be written

\[ i_{C1} = I_S e^{(v_{B1} - v_E) / V_T} \]  \hspace{1cm} (3.2)

\[ i_{C2} = I_S e^{(v_{B2} - v_E) / V_T} \]  \hspace{1cm} (3.3)

where $V_T = kT/q$. $k$ is Boltzmann's constant, $T$ is the temperature in °K, and $q$ is the electron charge.

These two equations can be combined to obtain

\[ \frac{i_{C1}}{i_{C2}} = e^{(v_{B1} - v_{B2}) / V_T} \]  \hspace{1cm} (3.4)

which can be manipulated to yield

\[ \frac{i_{C1}}{i_{C1} + i_{C2}} = \frac{1}{1 + e^{(v_{B2} - v_{B1}) / V_T}} \]  \hspace{1cm} (3.5)

\[ \frac{i_{C2}}{i_{C1} + i_{C2}} = \frac{1}{1 + e^{(v_{B1} - v_{B2}) / V_T}} \]  \hspace{1cm} (3.6)

For $\beta \gg 1$, $i_{C1} + i_{C2} = I_Q$ \hspace{1cm} (3.7)
Using Eq.(3.7) together with Eqs. (3.5) and (3.6) gives

\[ i_{C1} = \frac{I_Q}{1 + e^{(v_{B2} - v_{B1})/V_T}} \]  
(3.8)

\[ i_{C2} = \frac{I_Q}{1 + e^{(v_{B1} - v_{B2})/V_T}} \]  
(3.9)

The fundamental operation of the differential amplifier is illustrated by Eqs. (3.8) and (3.9). Note that the amplifier responds only to the difference voltage \( v_{B1} - v_{B2} \). That is, if \( v_{B1} = v_{B2} = v_{CM} \), the current \( I_Q \) divides equally between the two transistors independently of the value of the common-mode voltage \( v_{CM} \). This is the essence of differential-amplifier operation, which also gives rise to its name.
Fig. 3.8 The basic BJT differential-pair configuration.
3.3.3 The BJT Differential Amplifier with Active Loads

The application of a difference-mode input voltage \( v_i \), where \( v_i = v_{B1} - v_{B2} \), to a differential amplifier will produce ac collector currents of \( i_{c1} = -i_{c2} = g_m v_i \). Here \( g_m \) denotes the transconductance of \( Q_1 \) and of \( Q_2 \) \[14\]. To transform these ac currents into an output voltage, a load must be used. The load can take the form of a passive load, which uses a pair of load resistors \( R_L \) connected between the collectors of the differential amplifier and the dc supply as shown in Fig. 3.9, or an active load, which uses transistors for the current-to-voltage transformation (Fig. 3.10).

The simplest type of load is the passive load using load resistors \( R_L \). With this type of load, the ac output voltages at the two collectors are given by \( v_{c1} = -i_{c1} R_L = -g_m R_L v_i \) and \( v_{c2} = -i_{c2} R_L = g_m R_L v_i \), and the ac voltage gain for a single-ended output is \( A_v = -v_{c1} / v_i = v_{c2} / v_i = g_m R_L \). Thus, for a large voltage gain a large value of \( R_L \) is required. This can be seen more clearly by using the fact that \( g_m = I_Q / 4V_T \) for equal bias points, so that \( A_v = g_m R_L = I_Q R_L / 4V_T = I_Q R_L / 0.1V \) \[20\]. Since very small values of \( I_Q \) are often used in differential amplifiers (in the range of a few microamperes), we see that very large values of \( R_L \) will indeed be required (~1 MΩ) for substantial voltage gains\[27\]. However, the use of these large values of load resistance in IC’s carries with it some major disadvantages, such as taking up an excessive amount of room on the IC chip, limiting the frequency response, and reducing the input voltage range of the differential amplifier\[14\].

Most IC differential amplifiers use active loads, which involve the use of transistors rather than resistors. A simple active-load configuration for a differential amplifier is the current-mirror active load which has been shown in Fig. 3.10. Its operation is as follows: Transistors \( Q_1 \) and \( Q_2 \) form the differential pair biased with constant current \( I_Q \). The load
circuit consists of transistors Q3 and Q4 connected in a current mirror configuration. The output is taken single-endedly from the collector of Q2. Consider first the case when no input signal is applied (that is, the two input terminals are grounded.) The current $I_Q$ splits equally between $Q_1$ and $Q_2$. Thus $Q_1$ draws a current approximately $I_Q/2$ from the diode-connected transistor $Q_3$. Assuming $\beta \gg 1$, the mirror supplies an equal current $I_Q/2$ through the collector of $Q_4$. Since this current is equal to that through the collector of $Q_2$, no output current flows through the output terminal. Next consider the situation when a differential signal $v_d$ is applied at the input. Current signals $g_m v_d$ will result in the collectors of $Q_1$ and $Q_2$ with the polarities indicated in Fig. 3.10[14]. The current mirror reproduces the current signal $g_m v_d$ through the collector of $Q_4$. Thus, at the output node we have two current signals that add together to produce a total current signal of $(2g_m v_d)$. Now if the resistance presented by the subsequent amplifier stage is very large, the voltage signal at the output terminal will be determined by the total signal current $(2g_m v_d)$ and the total resistance between the output terminal and ground, $R_o$; that is, $v_o = 2g_m v_d R_o$. The output resistance $R_o$ is the parallel equivalent of the output resistance of $Q_2$ and the output resistance of $Q_4$. Since $Q_2$ is operating in the common-emitter configuration, its output resistance will be equal to $r_o$ of $Q_2$ - that is, $r_{o2} = V_A / I_{c2}$. The output resistance of the basic current mirror circuit is equal to $r_o$ of $Q_4$ - that is, $r_{o4} = V_A / I_{c4}$ [20], where $I_{c2}$ and $I_{c4}$ are the collector currents of transistors $Q_2$ and $Q_4$, respectively. Thus, $R_o = r_{o2} // r_{o4}$. 
Fig. 3.9 The differential amplifier with a passive load.
Fig. 3.10 The differential amplifier with an active load.
3.3.4 Biasing in BJT Integrated Circuits

With present BJT IC technology it is almost impossible to fabricate large capacitors, and it is uneconomical to manufacture large resistances. Instead, IC technology provides the designer with the possibility of using many transistors, which can be produced cheaply. Furthermore, it is easy to make transistors with matched characteristics that track with changes in environmental conditions. The limitations of, and opportunities available in, IC technology dictate a biasing philosophy that is quite different from that employed in discrete BJT amplifiers[26]. Basically, biasing in integrated-circuit design is based on the use of constant current sources. In the specific linear OTA circuit of this thesis, the differential pairs utilize constant current source bias. The configuration of the current sources is shown in Fig. 3.11. The transistors are scaled to have the emitter area factor shown. Otherwise, they are identical. The analysis proceeds as follows:

Since the relative device area of Q₁ is four times than one of Q₂, the emitter current of Q₁ is also four times than one of Q₂. This is the key point. The rest of the analysis is straightforward and is illustrated in Fig. 3.11.

This gives,

\[ I_{REF} = \frac{\beta(\beta + 1)}{(\beta + 1)^2} I_E + \frac{4\beta + 5}{(\beta + 1)^2} I_E \]  

\[ = \frac{\beta^2 + 5\beta + 5}{(\beta + 1)^2} I_E \]  

(3.10)

and

\[ I_o = \frac{4\beta^2 + 5\beta}{(\beta + 1)^2} I_E \]  

(3.11)
So,

\[
\frac{I_o}{I_{\text{REF}}} = \frac{4\beta^2 + 5\beta}{\beta^2 + 5\beta + 5}
\]

For \( \beta >> 1 \), \( I_o \equiv 4I_{\text{REF}} \).

In our OTA case, the schematic diagram of the constant-current sources generating \( I_{E1}, I_{E2}, I_{E3}, \) and \( I_{E4} \) of the OTA is shown in Fig. 3.12. The input common base stage is used to generate \( I_E \) because it is more convenient to connect an independent voltage source to the IC from outside. We use transistors \( Q_{E3} / Q_{E13} / Q_{E23} / Q_{E33} \) for the voltage-to-current transformation. By adjusting external independent voltage sources \( V_{A1}, V_{A2}, V_{A3}, \) and \( V_{A4} \), we can control \( I_{E1}, I_{E2}, I_{E3}, \) and \( I_{E4} \).
Fig. 3.11 A constant-current source.
Fig. 3.12 The schematic diagram of the constant-current sources $I_{E1}$, $I_{E2}$, $I_{E3}$, and $I_{E4}$ of the OTA.
3.3.5 Transconductance Amplifier

The transconductance amplifier of the OTA consists of two voltage-coupled differential pairs (Fig. 3.5). Transistors Q₁ and Q₂ in the inner pair are biased from the current source Iₑ₁ and work with a local series feedback formed by the resistors R₁, R₁₁, capacitors C₁, C₁₁ and diode-connected transistors Q₃ and Q₄. Transistors Q₅ and Q₆ in the outer pair are biased from the separate current source Iₑ₂. The transconductance gain of the OTA, $g_{mT}$, is found as follows:

The $i_c-\Delta v_{be}$ relationship of a forward-biased transistor is $i_c = I_S \exp(\Delta v_{be}/V_T)$, where $I_S$ is a constant and $V_T = kT/q$ is the thermal voltage.

Let $R_1 = R_{11}$, $C_1 = C_{11}$, as shown in Fig. 3.13. By KVL,

$$\Delta v_{in} = \Delta v_{be1} + Z_1 \Delta i_1 - \Delta v_{be3} - \Delta v_{be4} + Z_1 \Delta i_1 - \Delta v_{be2} \quad (3.13)$$

where

$$Z_1 = \frac{1}{C_1 S + \frac{1}{R_1}} = \frac{R_1}{R_1 C_1 S + 1}$$

Because the same collector current biases each transistor, we can assume the transconductance gain of each transistor is same. i.e. $g_{m1} = g_{m2} = g_{m3} = g_{m4} = g_M$

where

$$g_{mi} = \frac{\Delta i_{ci}}{\Delta v_{be1}}$$

$i = 1, 2, 3, 4$, is the definition of transconductance of each individual transistor.
Since the betas are large for these devices, $\Delta i_1 = \Delta i_c$, and:

$$\Delta v_{in} = \frac{\Delta i_c}{g_M} + Z_1 \Delta i_c + \frac{\Delta i_c}{g_M} - \frac{\Delta i_c}{g_M} + Z_1 \Delta i_c - \frac{\Delta i_c}{g_M}$$

This gives

$$\Delta v_{in} = 4 \frac{\Delta i_c}{g_M} + 2Z_1 \Delta i_c$$

So

$$\frac{\Delta i_c}{\Delta v_{in}} = 4 \frac{1}{2(Z_1 + \frac{2}{g_M})} = \frac{g_M}{2(2 + g_M Z_1)}$$

If $g_M Z_1 >> 1$, the transconductance of the input stage is:

$$G_{m1} = \frac{\partial i_c}{\partial v_{in}} = \frac{g_{m1}}{2(2 + g_{m1} Z_1)} = \frac{1}{2Z_1}$$

The transconductance $g_{m1}$ of transistor $Q_1$ (or $Q_2$) can be directly related to its bias current as

$$g_{m1} = \frac{\partial i_c}{\partial v_{be1}} = \frac{I_{E1}}{2V_T} = \frac{I_{E1}}{2V_T}$$

where $I_{E1}$ is the bias current to the transistor pair (Fig. 3.14).
From Eq. (3.16), \( G_{m1} \) is approximately constant and no electronic adjusting of the first stage gain is possible. In order to implement this adjustment of the transconductance, a second stage is introduced. As seen from Fig. 3.5, transistor Q3 and Q5 have different bias currents, but there is a signal path from Q3 to Q5. The transconductance of the second stage is equal to that of Q5 and can be written as

\[
\begin{align*}
G_{m5} &= \frac{\partial i_{c5}}{\partial v_{be5}} = \frac{i_{c5}}{v_T} = \frac{i_{E2}}{2v_T} \\
\end{align*}
\]

For the second stage differential amplifier Q5 and Q6, as shown in Fig. 3.14, the common-mode input voltage \( v_{cm2} = \frac{(v_{cm1} - v_{be3}) + (v_{cm1} + v_{be3})}{2} = v_{cm1} \) and the differential-mode input voltage \( v_{dm2} = v_{be3} + v_{cm1} - (v_{cm1} - v_{be3}) = 2v_{be3} \). Note that, for the differential amplifier, the differential-mode voltage \( v_{dm2} = 2v_{be} \) is amplified, but the common-mode voltage \( v_{cm2} = v_{cm1} \) is not. The transconductance of this balanced differential amplifier is

\[
G_{m2} = \frac{i_{E2}}{4v_T}
\]

Therefore,

\[
\Delta i_{c5} = G_{m2}\Delta(2v_{be3}) = G_{m2}\frac{\Delta i_{c3}}{G_{m3}} = G_{m2}\frac{\Delta i_{c3}}{G_{m1}}
\]

but \( \Delta i_{c3} = G_{m1}\Delta v_{in} \)
Define the amplifier’s transconductance as:

\[ g_{mT} = \frac{\Delta i_c}{\Delta v_{in}} = -\frac{\Delta i_c}{\Delta v_{in}} \]

So,

\[ g_{mT} = \frac{\Delta i_c}{\Delta v_{in}} = G_m2 \frac{2}{G_{m1}} G_{m1} = (\frac{2}{2Z_1})(\frac{I_{E2}}{4V_T})(\frac{2V_T}{I_{E1}}) = \frac{I_{E2}}{2I_{E1}Z_1} \]  

(3.21)

For low frequency, the transconductance of the OTA is:

\[ g_{mT} = \frac{I_{E2}}{2I_{E1}R_1} \]  

(3.22)

which indicates that either \( R_1 \) or the ratio of \( I_{E1} \) and \( I_{E2} \) can be used for setting \( g_{mT} \).

The BJT model of the npn HBT device has a low Early voltage ( \( V_A = 20 \), see Table 3.1 ). This results in relatively low output resistances for transistors Q5, Q6 and Q9, Q10 ( because \( r_o = V_A / I_C \) ). To increase the output impedance of the OTA, a cascode output stage with transistors Q7, Q8 is used[20].

If the base current Q7 is neglected, the total transconductance \( g_{mT} \) is unchanged by this stage:

\[ g_{mT} = \frac{\partial i_c}{\partial v_{in}} \approx \frac{\partial i_c}{\partial v_{in}} \]  

(3.23)
Fig. 3.13  The ac circuit analysis of the transconductance of the input stage of the OTA.
Fig. 3.14 The ac circuit analysis of the transconductance of the second stage of the OTA.
3.3.6 The Common-Mode Feedback Circuit

One of the basic problems of fully-differential operation of an OTA is that the feedback around the amplifier only provides a Differential-Mode (DM) feedback, and therefore a stabilization of the quiescent DM signal levels. The same feedback does however not provide any stabilization of the common-mode (CM) signal levels. Therefore, CM output is undefined and the transistors in the output stage may drift out of their linear range. To avoid this problem, an extra common-mode feedback (CMF) loop must be employed. The simplified block diagram of a fully-differential transconductance amplifier with Common-Mode feedback is shown in Fig. 3.15. The CMF circuit senses the CM output voltage $V_{o,CM}$ by a circuit which is ideally insensitive to the DM output voltage $V_{o,DM}$, where $V_{o,CM} = (V_{o,+} + V_{o,-})/2$ and $V_{o,DM} = V_{o,+} - V_{o,-}$. The output signal of the CMF circuit is fed back to the amplifier to adjust the DC current from each output driver so that $V_{o,CM}$ can be maintained at a constant DC level, normally at zero.

The schematic diagram of the specific CMF circuit in the OTA has been shown in Fig. 3.6. Each of two differential pairs ($Q_{11}, Q_{13}$) and ($Q_{12}, Q_{14}$) senses one of the output voltages and compares it with the ground. The outputs of the two pairs are tied together in such a way that the differential signals cancel and the common mode signal is amplified. Transistors $Q_{15}$ and $Q_{16}$ form the active load of the two differential pairs ($Q_{11}, Q_{13}$) and ($Q_{12}, Q_{14}$).

Fig. 3.16 shows the circuit analysis of the CMF circuit used in the simulated OTA. Fig. 3.17 shows the small-signal model of the BJT, used in this analysis. Here the amplifier is working open loop, that is, the base of $Q_{11}$ (node 160) is normally connected to node 16, and node 120 is normally part of node 12, but the signal paths have been
broken to examine the loop gain. By circuit analysis, we see that there is a strong negative feedback through the amplifier to the output nodes.

For a common mode input signal, $v_{cm} = (v_{160} + v_{120}) / 2$, the output voltages are:

$$v_{16} = -\beta_0 \alpha_n R_{L1} \left( \frac{v_{cm}}{r_e + R_E} \right)$$ (3.24)

$$v_{12} = -\beta_0 \alpha_n R_{L2} \left( \frac{v_{cm}}{r_e + R_E} \right)$$ (3.25)

So, the common mode output voltage is:

$$v_{cmout} = \frac{v_{16} + v_{12}}{2} = -\beta_0 \alpha_n v_{cm} \left( R_{L1} + R_{L2} / (r_e + R_E) \right)$$ (3.26)

Fig. 3.18 shows an equivalent OTA circuit which includes the effect of common mode feedback. The output voltages are:

$$v_{16} = (-T v_{cm} + g_m v_{in}) R_{L1}$$ (3.27)

$$v_{12} = (-T v_{cm} - g_m v_{in}) R_{L2}$$ (3.28)

where

$$T = -\frac{\beta_0 \alpha_n}{r_e + R_E}$$

is the open loop common mode feedback transconductance gain.
So, the common mode output voltage is:

\[
V_{cmout} = \frac{v_{16} + v_{12}}{2} = -\frac{T V_{cm}(R_{L1} + R_{L2})}{2} + \frac{g_m v_{in}(R_{L1} - R_{L2})}{2}
\]

(3.29)

This gives

\[
V_{cmout} = \left[\frac{1}{1 + \frac{T(R_{L1} + R_{L2})}{2}}\right] \frac{g_m v_{in}(R_{L1} - R_{L2})}{2}
\]

(3.30)

If \(R_{L1} = R_{L2}\), then \(V_{cmout} = 0\).

If \(R_{L1} \neq R_{L2}\) and the gain \(T\) is large, then \(V_{cmout}\) is still small.

Fig. 3.19 shows a modified OTA circuit including the input amplifier. The connection has been broken between the transconductance amplifier and the CMF circuit again in order to investigate the actual common mode feedback open loop gain with the original transconductance amplifier in place. The result of the PSPICE simulations on a transistor level is shown in Fig. 3.20. The PSPICE input circuit file is included in Appendix B. From the results, we see that the actual common mode feedback open loop gain of the CMF circuit is a complex number and is dependent on frequency. Here, at the frequency of 1 GHz, the magnitude of the common mode feedback gain of the common mode amplifier (CMA) is \(15.6 \times 10^{-3}\) and the phase of the gain is 43.8 degrees.

By Eqs. (3.27) and (3.28), \(-T \cdot R_{L1} = 15.6 \cdot 10^{-3} \leq 43.8^\circ\).
This gives $T = -7.8 \cdot 10^{-4} \angle 43.8^\circ$. Here, $R_{L1} = 20$ ohms and $v_{in} = 0$.

We see that the open loop gain $T$ is quite small in the OTA. Therefore, by Eq. (3.30), $R_{L1}$ and $R_{L2}$ are needed to be equal in order to keep

$$v_{cm_{out}} = 0.$$
Fig. 3.15 The simplified block diagram of a fully-differential transconductance amplifier with common-mode feedback.
Fig. 3.16 Circuit analysis of the CMF circuit (open loop).
Fig. 3.17 The small-signal model of the BJT used in the analysis of Fig. 3.16.
Fig. 3.18  An equivalent OTA circuit which includes the first order effect of the common mode feedback.
Fig. 3.19

A modified OTA circuit for investigation of common mode open loop feedback gain.
Fig. 3.20  
(a) Common mode feedback gain of the CMA. 
(b) Phase of common mode feedback gain of the CMA. 
(The common mode input signal $V_{IN1}=V_{IN2}=1mV, R_{L1}=R_{L2}=20$ ohms)
3.4 PSPICE Simulations of the OTA on the Device Level

Table 3.2 shows the PSPICE parameter values of components of the specific OTA circuit considered in this thesis. The test circuit used in simulation of the OTA circuit is presented in Fig. 3.21. The PSPICE input circuit file is included in Appendix A. DC and AC analyses from the PSPICE simulation of the OTA for $R_{L1}=R_{L2}=50$ ohms on the transistor level are shown in Fig. 3.22. Fig. 3.22 (a) shows the two output DC signal curves of the OTA are symmetric. Fig. 3.22 (b) shows the input range of the OTA which gives a linear output is about ±1.5V. Fig. 3.22 (c) shows that the low frequency transconductance gain of the OTA, $g_{mT}$, is 2.48 millimhos and the upper 3dB cut-off frequency is around 42 GHz. Fig. 3.22 (d) shows the phase of the frequency response of $g_{mT}$. The results of PSPICE simulations for changing $g_{mT}$ are presented in Fig. 3.23 - Fig. 3.25. Fig. 3.23 (e) shows that $g_{mT}$ seems to go as $1/R_{1}$. It is shown in Fig. 3.25 (e) that $g_{mT}$ is proportional to $I_{E2}$. In Fig. 3.24 (e), we observe that $g_{mT}$ goes as $(I_{E1})^{-1/2}$. (The standard theory based on the Ebers-Moll equations predicted the $g_{mT}$ should go as $(I_{E1})^{-1}$ as was demonstrated in Eq. (3.23). However, PSPICE uses a more advanced model for a transistor, namely the Gummel-Poon model[34, 37]. This may explain the difference).

Thus, it is possible to change $g_{mT}$ by either changing $I_{E1}$ or $I_{E2}$. In addition, the effect of the output load resistors $R_{L1}$ and $R_{L2}$ on the DC and AC responses of this OTA are shown in Fig. 3.26. The output load resistors $R_{L1}$ and $R_{L2}$ will affect the linear input range and the position of the upper 3dB cut-off frequency very seriously, but not the low frequency transconductance gain $g_{mT}$. The larger the output load resistors, the narrower the linear input range and the smaller the upper 3dB cut-off frequency.
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<th>Parameter of Component</th>
<th>Values</th>
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<tr>
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Table 3.2  The PSPICE parameter values of components of the OTA circuit.
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Table 3.2  The PSPICE parameter values of components of the OTA circuit (continued).
Table 3.2 The PSPICE parameter values of components of the OTA circuit (continued).
Fig. 3.21 The test circuit used in simulation of the OTA circuit
Fig. 3.22  (a) DC analysis of the OTA.
(b) Slope of the DC curve showing the linear range.
\((R_{L1}=R_{L2}=50 \text{ ohms})\)
(c) Frequency response of the transconductance of the OTA.
(d) Phase of the transconductance ($g_{mT}$).
($R_{L1}=R_{L2}=50$ ohms)
Fig. 3.23  
(a) DC analysis of the specific OTA as resistors $R_1$ and $R_{11}$ are changed.  
(b) Slope of the DC curves in order to recognize the linear range easily.  
($I_{E1} = 8.2\, \text{mA}$, $I_{E2} = 10.1\, \text{mA}$, $R_{L1}=R_{L2}=50\, \text{ohms}$)
Fig. 3.23  
(c) Frequency response of output current for different values of $R_1$ and $R_{11}$.  
(d) Phase of the frequency response of output current for different values of $R_1$ and $R_{11}$.  
($I_{E1} = 8.2\text{mA}, I_{E2} = 10.1\text{mA}, R_{L1}=R_{L2}=50\text{ ohms}$)
Fig. 3.23  (e) The relationship between $g_{mT}$ and $R_1$. 

$$g_{mT} \propto (R_1)^{-0.89}$$
Fig. 3.24  
(a) DC analysis of the OTA for changing bias current $I_{E1}$.  
(b) Slope of the DC curves showing the linear range. 
($R_1=R_{11}=180$ ohms, $I_{E2} = 10.1$ mA, $R_{L1}=R_{L2}=50$ ohms)
Fig. 3.24  (c) Frequency response of the output ac current as bias current $I_{E1}$ changes.
(d) Phase of the frequency response.
($R_1=R_{11}=180$ ohms, $I_{E2} = 10.1$ mA, $R_{L1}=R_{L2}=50$ ohms)
Fig. 3.24 (e) The relationship between $g_mT$ and $I_{E1}$.

$g_mT \propto (I_{E1})^{-0.54}$
Fig. 3.25  (a) DC analysis of the OTA as bias current $I_{E2}$ is changed. (b) Slope of the DC curves showing the linear range. ($R_1= R_{11} = 180$ ohms, $I_{E1} = 8.2$ mA, $R_{L1} = R_{L2} = 50$ ohms)
Fig. 3.25  
(c) Frequency response of the output current as bias current $I_{E2}$ changes.
(d) Phase of the frequency response.
($R_1=R_{11}=180$ ohms, $I_{E1} = 8.2$ mA, $R_{L1}=R_{L2}=50$ ohms)
Fig. 3.25 (e) The relationship between $g_{mT}$ and $I_{E2}$. 
Fig. 3.26  (a) DC analysis of the OTA for different load resistors (R_{L1} and R_{L2}).
(b) Slope of the DC curves to show the linear range.
Fig. 3.26  
(c) Frequency response of the OTA for different load resistors ($R_{L1}$ and $R_{L2}$).  
(d) Phase of the frequency response of the OTA for different load resistors ($R_{L1}$ and $R_{L2}$).
3.5 Summary

In this chapter we have used GaAs HBTs in a $g_m$ adjustable high-frequency fully differential OTA. The transconductance $g_{mT}$ of the OTA can be changed by setting some external voltages and a pair of resistors connected to the OTA. Simulations show the amplifier is broadband from DC to 42GHz, the maximum linear input range is $\pm 1.5$V and the transconductance coefficient $g_{mT}$ has values up to 2.48 millimhos. The output load resistors of the OTA affect the linear input range and the position of the upper 3dB cut-off frequency. A built-in feedback amplifier minimizes the common mode output voltage, which provides stabilization of the common mode signal levels of the OTA.
CHAPTER 4

BIQUAD TRANSCONDUCTANCE-RC (gm-RC) BANDPASS
ACTIVE FILTER

4.1 Introduction

Filters are electrical networks that process signals on a frequency-dependent basis and are typically categorized according to the filtering function performed. If magnitude (gain, attenuation) requirements are of primary importance, the filters are classified as lowpass, highpass, bandpass, or bandreject network. If, on the other hand, we are mainly concerned with phase or delay specifications with no change in magnitude, the filters typically will be allpass networks or delay equalizers[20]. If filters employ only passive components such as resistors, capacitors, and inductors, we name them as passive filters. If filters use transistors, operational amplifiers, or operational transconductance amplifiers together with passive components, they are active filters[25].

In this chapter, we will discuss the basic characteristics of second-order filter functions and introduce a specific biquad transconductance-RC (gm-RC) bandpass active filter utilizing three OTAs. We want to do a simple model of the OTA to use, but need to do PSPICE analysis to check on the validity of the model. Therefore, the results of PSPICE simulations on the ideal OTA model level and the transistor level are shown.
4.2 Basic Characteristics of Second-Order Filter Functions

The way in which a filter function varies with frequency is called frequency response. This can be represented mathematically by means of the filter's transfer function $H$. Filter approximation is most conveniently performed using Laplace transform theory. The result of the approximation process is a function $H(s)$ from which we can obtain the frequency response function by substituting $s = j\omega = j2\pi f$.

In any type of filter, the coefficients of the transfer function

$$H(s) = \frac{N(s)}{D(s)} = \frac{a_m s^m + \ldots + a_1 s + a_0}{s^n + \ldots + b_1 s + b_0}$$

have to be determined such that the desired filter specifications are met. Here $n$ is the order of the filter. For stability reasons, the degree of the numerator polynomial must be less than or equal to that of the denominator polynomial, that is, $m \leq n$ [25]. The numerator and denominator coefficients, $a_0, a_1, ..., a_m$ and $b_0, b_1, ..., b_{n-1}$, are real numbers. The numerator and denominator polynomials can be factored and $H(s)$ can be expressed in the form

$$H(s) = \frac{a_m(s - z_1)(s - z_2) \ldots (s - z_m)}{(s - p_1)(s - p_2) \ldots (s - p_n)}$$

Here the numerator roots, $z_1, z_2, ..., z_m$, are called the transfer function zeros, and the denominator roots, $p_1, p_2, ..., p_n$, are called the transfer function poles. Each zero or pole can be either a real or a complex number. However, complex zeros and poles must occur in conjugate pairs since the polynomial coefficients are real.
A particularly important class of filter functions realizes the second-order transfer function. A second-order filter function can be written as the ratio of two quadratic polynomials,

\[ H(s) = \frac{\mathcal{N}(s)}{\mathcal{D}(s)} = \frac{a_2 s^2 + a_1 s + a_0}{s^2 + b_1 s + b_0} = \frac{a_2 (s - z_1) (s - z_2)}{(s - p_1) (s - p_2)} \]

and thus is known as a biquadratic function or simple a biquad. It serves as the building block for a wide variety of active filters\[25\].

For complex poles and zeros, where \( z_2 = z_1^* \) and \( p_2 = p_1^* \)

( the superscript * denotes a complex conjugate ), we can express Eq. (4.3) as

\[ H(s) = \frac{S^2 - [2\text{Re}(z_1)] S + \text{Re}(z_1)^2 + \text{Im}(z_1)^2}{S^2 - [2\text{Re}(p_1)] S + \text{Re}(p_1)^2 + \text{Im}(p_1)^2} \]  

\[ = \frac{\frac{\omega_z^2}{Q_z} S + \omega_z^2}{\frac{\omega_p^2}{Q_p} S + \omega_p^2} \]

where \( \text{Re}(z) \) and \( \text{Im}(z) \) denote the real part and the imaginary part of \( z \), respectively.

Equation (4.4b) is the standard notation for biquadratic functions because it identifies clearly the important filter performance parameters. The dc gain and the asymptotic gain as \( \omega \to \infty \) are given by
and

\[ 20 \log_{10} |H(j0)| = 20 \log_{10}(K \frac{\omega_p^2}{\omega_0^2}) \]

and

\[ 20 \log_{10} |H(j\infty)| = 20 \log_{10}(K) \]

The sharpness of the maximum or bump at \( \omega_p \) is determined by the pole quality factor \( Q_p \), defined as

\[
Q_p = \frac{\omega_p}{2 \text{Re}(p_1)} = \frac{\sqrt{\text{Re}(p_1)^2 + \text{Im}(p_1)^2}}{2 \text{Re}(p_1)}
\]

and the depth of the minimum at \( s = j \omega_z \) is determined by the zero quality factor

\[
Q_z = \frac{\omega_z}{2 \text{Re}(z_1)} = \frac{\sqrt{\text{Re}(z_1)^2 + \text{Im}(z_1)^2}}{2 \text{Re}(z_1)}
\]

A special case of the general biquadratic function [ Eq. (4.3) ], which we will consider in section 4.3, is the bandpass (BP) function. In Eq. (4.3), if \( a_0 = a_2 = 0 \), we obtain this function, denoted by

\[
H_{BP}(s) = \frac{a_1 s}{s^2 + s(\frac{\omega_0}{Q}) + \omega_0^2}
\]
Thus there is one zero at \( s = 0 \) (dc) and one at \( s = \infty \). Fig. 4.1 shows the magnitude response of a second order bandpass filter with unity center-frequency gain (\( a_1 = \omega_0 / Q \)).

Note that the response peaks at \( \omega = \omega_0 \); thus \( \omega_0 \) is also called the center frequency. The pole quality factor \( Q \) determines how narrow (selective) the bandpass filter is. Specifically, the two frequencies \( \omega_1 \) and \( \omega_2 \) at which the transmission drops by 3dB below the value at \( \omega_0 \) are separated by \( \omega_0 / Q \). This distance is called the 3dB bandwidth. As \( Q \) is increased, the bandwidth decreases and the filter becomes more selective.
Fig. 4.1  The magnitude response of a second order bandpass filter with unity center-frequency gain.
4.3 A Biquad $g_m$-RC Bandpass Active Filter with a Simple OTA Model

The schematic diagram of the practical biquad $g_m$-RC bandpass active filter is shown in Fig. 4.2. First, we use the ideal OTA model which is a voltage-controlled current source whose input and output impedance are both infinite. Fig. 4.3 shows the biquad $g_m$-RC bandpass active filter with the simple OTA model. With circuit analysis, the transfer function of the circuit can be written as:

\[
\frac{V_{\text{out}}}{V_{\text{in}}} = \frac{sC_1g_mT_3}{s^2C_1C_2 + sC_1g_mT_2g_mT_3R_1 + g_mT_2g_mT_3}
\] (4.8)

By comparing to Eq. (4.7), one sees the center frequency and pole quality factor are

\[
\omega_0^2 = \frac{g_mT_1g_mT_3}{C_1C_2}
\] (4.9)

and

\[
Q = \frac{\omega_0C_2}{g_mT_2g_mT_3R_1}
\] (4.10)

In order to have a comparison with the filter on the transistor level later, we assume that $g_mT_1 = g_mT_2 = g_mT_3 = 2.48$ millimhos in the filter and choose $R_1=60$ ohms, $C_1=1$ pF, and $C_2=0.01$ pF. By Eq. (4.9) and (4.10), we get $f_0 = \omega_0/(2\pi) = 3.95$ GHz and $Q = 0.67$. 
The PSPICE simulation is shown in Fig. 4.4. The center frequency $f_0$ is 3.98 GHz and the quality factor $Q$ is 0.68. The results agree with the equations above. The phase of the bandpass filter shown in Fig. 4.4 (b) is smooth and well behaved (zero degrees and monotonically decreasing) around the center frequency. This means the bandpass filter circuit is stable. Fig. 4.5 shows the effect of resistor $R_1$ on the frequency response, while Fig. 4.6 shows the effect of capacitor $C_1$. The effect of capacitor $C_2$ is shown in Fig. 4.7.
Fig. 4.2 The schematic diagram of a biquad $g_m$-RC bandpass active filter.
Fig. 4.3 The $g_m$-RC bandpass filter utilizing the ideal OTA model in Fig. 4.2.
Fig. 4.4 (a) The magnitude and (b) the phase of the gain of the biquad $g_m$-RC bandpass filter utilizing a simple OTA model. ($g_{mT1}=g_{mT2}=g_{mT3}=2.48$ millimhos, $C_1=1$ pF, $C_2=0.01$ pF, $R_1=60$ ohms).
Fig. 4.5 Effect of resistor $R_1$ on (a) the magnitude and (b) the phase of the voltage gain of the biquad $g_m$-RC bandpass active filter utilizing the simple OTA model. ($g_{mT1}=g_{mT2}=g_{mT3}=2.48$ millimhos, $C_1=1$ pF, $C_2=0.01$ pF).
Fig. 4.6 Effect of capacitor $C_1$ on (a) the magnitude and (b) the phase of the voltage gain of the biquad $g_m$-RC bandpass active filter utilizing the simple OTA model. ($g_{m1}=g_{m2}=g_{m3}=2.48$ millimhos, $R_1=60$ ohms, $C_2=0.01$ pF).
Fig. 4.7 Effect of capacitor $C_2$ on (a) the magnitude and (b) the phase of the voltage gain of the biquad $g_m$-RC bandpass active filter utilizing the simple OTA model. ($g_{mT1}=g_{mT2}=g_{mT3}=2.48$ millimhos, $R_1=60$ ohms, $C_1=1$ pF).
4.4 The Results of PSPICE Simulations on the $g_m$-RC Bandpass Active Filter on the Transistor Level

The schematic diagram of a specific biquad $g_m$-RC bandpass active filter utilizing three OTAs which we have introduced in chapter 3 is presented in Fig. 4.8. In order to keep the common mode output voltage on each OTA small, the load resistors $R_{L1}$, $R_{L2}$, and $R_{L3}$ are needed in an actual filter (refer to the discussion in section 3.3.6). Table 4.1 shows the parameter values of the components used in the filter. The internal values of the OTAs have been shown in Table 3.2.

PSPICE simulations show the values of output terminal resistors $R_{L1}$, $R_{L2}$, and $R_{L3}$ affect the gain, the quality factor $Q$, and the center frequency of the bandpass filter (see Fig. 4.9). Therefore, by changing the OTA output terminal resistors $R_L$, the $Q$ can be controlled. This is simpler than using a very complex $Q$-controlled circuit[1, 25]. The phases of the bandpass filter shown in Fig. 4.9 (b) are smooth and well behaved (zero degrees and monotonically decreasing) around the center frequencies. So, the filter is stable. Fig. 4.10 shows the best result which has $Q = 139$ and $f_0 = 1.4$ GHz (after setting each OTA output load to $R_{L1} = R_{L2} = R_{L3} = 350$ ohms). Fig. 4.11 shows the output voltage of the filter as the resistor $R_1$ is varied. Fig. 4.12 shows the output voltage of the filter for different values of $C_1$. Fig. 4.13 shows the output voltage of the filter at several values of $C_2$. The PSPICE input circuit file is included in Appendix C.

In comparing Fig. 4.5 and Fig. 4.9, we see that the filter simulated on the transistor level has the phenomena of a high-$Q$ and that the $Q$ is changed by the OTA output load $R_L$. Because an exact representation of the electronic OTA circuit would be too complicated to handle and analyze without good software tools, it is desirable to develop an approximately model, but one that contains all relevant OTA imperfections with sufficient accuracy to permit realistic prediction of OTA and active filter behavior. Such a model is discussed in detail in the next chapter.
Fig. 4.8 The schematic diagram of the practical biquad $g_m$-RC bandpass filter.
<table>
<thead>
<tr>
<th>Devices</th>
<th>Parameter of Component</th>
<th>Values</th>
</tr>
</thead>
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<tr>
<td>Resistor</td>
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</tr>
<tr>
<td>&quot;</td>
<td>RL2</td>
<td>350 ohms</td>
</tr>
<tr>
<td>&quot;</td>
<td>RL3</td>
<td>350 ohms</td>
</tr>
<tr>
<td>&quot;</td>
<td>R1</td>
<td>60 ohms</td>
</tr>
<tr>
<td>Capacitor</td>
<td>C1</td>
<td>1 PF</td>
</tr>
<tr>
<td>&quot;</td>
<td>C2</td>
<td>0.01 PF</td>
</tr>
</tbody>
</table>

Table 4.1 Component values used in the bandpass filter.
Fig. 4.9  Effect of output terminal resistors $R_{L1}$, $R_{L2}$, and $R_{L3}$ on (a) the magnitude and (b) the phase of output voltage gain of the biquad $g_m$-RC bandpass filter simulated on the transistor level. ($R_1=60$ ohms, $C_1=1\mu F$, $C_2=0.01\mu F$)
Fig. 4.10 (a) The magnitude and (b) the phase of the output voltage gain of the biquad $g_m$-RC bandpass active filter as simulated on the transistor level ($R_{L1} = R_{L2} = R_{L3} = 350$ ohms).
Fig. 4.11 (a) The magnitude and (b) the phase of the output voltage gain of the biquad $g_m$-RC bandpass filter as resistor $R_1$ is varied. Simulated on the transistor level.

($R_{L1}=R_{L2}=R_{L3}=350 \text{ ohms}, C_1=1\mu\text{F}, C_2=0.01\mu\text{F}$)
Fig. 4.12  (a) The magnitude and (b) the phase of the output voltage gain of the biquad $g_m$-RC bandpass filter as capacitor $C_1$ is varied.
Simulated on the transistor level.
($R_{L1}=R_{L2}=R_{L3}=350$ ohms, $R_1=60$ ohms, $C_2=0.01$ pF)
Fig. 4.13  (a) The magnitude and (b) the phase of the output voltage gain of the biquad $g_m$-RC bandpass filter as capacitor $C_2$ is varied. Simulated on the transistor level.

(R$L_1$=R$L_2$=R$L_3$=350 ohms, $R_1$=60 ohms, $C_1$=1 pF)
CHAPTER 5

SIMULATION OF THE BANDPASS FILTER USING AN OTA MODEL

5.1 Introduction

As discussed in chapter 4, the filter has a high-Q characteristic and has the ability of the OTA output load $R_L$ to change Q. In order to provide for accurate design verification and to implement a procedure for accurate filter design, we need a more realistic OTA model. To get more realistic results than the simple current generator gives, a number of non-ideal OTA characteristics which may adversely affect the performance of the bandpass filter need to be incorporated into the model, so that we may accurately predict the center frequency and the Q of the filter.

In this chapter, we discuss non-ideal OTA characteristics and propose an OTA model which includes non-ideal characteristics. The use of this model of the OTA in the filter is explored and illustrated. The results of PSPICE simulations are shown.
5.2 Nonidealities of the OTA

A. Output terminal resistor $R_L$

The first nonideality of the OTA to be considered is the need for an output terminal resistor $R_L$ between the output and ground. As discussed in section 3.3.6, because of the common mode feedback amplifier in the OTA, the output terminal resistor $R_L$ is needed to keep the common mode output small (refer to Eq. (3.31)). This stabilizes the common-mode (CM) signal levels of the OTA's output.

B. Frequency dependent $g_{mT}$

The actual transconductance gain $g_{mT}$ of the OTA circuit is a frequency dependent function. The frequency dependence of $g_{mT}$ can be modeled via a dominant pole[2]

$$g_{mT}(j\omega) = \frac{g_{mo}}{1 + j\frac{\omega}{\omega_m}} = g_{ma}e^{-j\Phi(\omega)}$$  \hspace{1cm} (5.1)

where

$$\Phi(\omega) = \tan^{-1}\left(\frac{\omega}{\omega_m}\right)$$  \hspace{1cm} (5.2)

$$g_{ma} = \frac{g_{mo}}{\sqrt{1+\left(\frac{\omega}{\omega_m}\right)^2}}$$  \hspace{1cm} (5.3)

and $\omega_m$ is the -3dB radian frequency of the OTA circuit.
In our case, PSPICE analysis shows the -3dB frequency $f_m$ of the OTA circuit is 42GHz and the center frequency $f_0$ of the biquad $g_{m}-RC$ bandpass active filter is 1.15GHz (refer to Fig. 3.22 (c) and Fig. 4.10). Therefore, $f_m \gg f_0$. From Eq. (5.1), we get $g_{mT}(j\omega) = g_{m0} = \text{constant}$ when $\omega = 2\pi f_0$ is the working frequency of the OTA, which is around the center frequency $f_0$ of the filter. So, we can assume that the transconductance gain $g_{mT}$ is a constant in a realistic OTA model which can describe the behavior of this filter. However, if $f_m$ is not very much larger than $f_0$, then $g_{mT}(j\omega)$ is a frequency dependent function.

C. Finite input and output impedances

In practice, OTAs are electronic circuits, composed of many transistors, resistors, and capacitors. Consequently, their input and output impedance are not infinite. The input and output impedances may be assumed to model not only the parasitic device input and output impedances but also those contributed by layout and routing. For low-frequency applications the parasitic capacitances of OTAs are much smaller than the external capacitors used in the filter, but for high-frequency applications, the latter are forced to be small to get high frequency poles. It is necessary to consider parasitic capacitances of the devices that are no longer negligible compared with the external capacitors needed for filter design.
5.3 The Proposed OTA Model

A realistic OTA model which includes the nonidealities of an OTA is shown in Fig. 5.1. Note that because the OTA is completely symmetric, that is $R_1 = R_{11}$, $R_9 = R_{19}$ etc (refer to Table 3.2), the model is symmetric. PSPICE and FORTRAN programs were used to extract the parameters of this OTA model. The extraction method is presented below:

A. Extraction of Input Common-Mode Impedance.

The extraction circuit structure is shown as Fig. 5.2 (a). The inputs are set at $V_{in1} = V_{in2} = 1mV$ and the outputs are terminated in resistors $R_{L1} = R_{L2} = 350$ ohms. PSPICE was used to extract the current into terminal $1$, say $I_1$.

A consideration of fig. 5.2 (b) shows the input common-mode impedance of the OTA is

$$Z_{icm} = \frac{V_{in1}}{I_1} \quad \ldots \quad (5.4)$$

The phase of $Z_{icm}$ indicates it has a capacitive component, so:

$$Z_{icm} = R_{icm} \parallel \frac{1}{sC_{icm}} \quad \ldots \quad (5.5)$$

A FORTRAN program was used for this case and all the following cases to do the translation work of finding the impedances at different frequencies (refer to APPENDIX E - J for these programs).
B. Extraction of the Input Differential-Mode Impedance of the OTA.

The extraction circuit is shown in Fig. 5.2 (a). Refer to Fig. 5.3, let \( V_{in1}=1\text{mV} \), \( V_{in2}=-1\text{mV} \) and \( R_{L1}=R_{L2}=350 \text{ ohms} \). Use PSPICE on the transistor level to extract the current into node 1, say \( I \). The current which passes through the input differential mode impedance, say \( I_2 \), is \( I_2 = I - I_1 \), where \( I_1 \) has been defined in part A.

Therefore, the input differential-mode impedance of the OTA is

\[
Z_{idm} = \frac{V_{in1} - V_{in2}}{I_2}
\]

The results show that \( Z_{idm} \) has a capacitive part, so:

\[
Z_{idm} = R_{idm} // \frac{1}{sC_{idm}}
\]

C. Extraction of Input Impedance of the OTA.

The extraction circuit structure is again shown as Fig. 5.2 (a). This time, \( V_{in1}=1\text{mV} \), \( V_{in2}=-1\text{mV} \). The RL’s are unchanged. PSPICE was used to extract the current into node 1 again, say \( I_3 \).

The input impedance of the OTA is

\[
Z_{in} = \frac{V_{in1} - V_{in2}}{I_3}
\]
D. The Output Common-Mode Impedance.

The extraction circuit structure is shown as Fig. 5.4 (a). Here, $V_{in3}=V_{in4}=1\text{mV}$ and $RL1=RL2=350$ ohms. Note that $RL1$ and $RL2$ are needed to keep the common mode amplifier from being overdriven. PSPICE was used to find the current into the terminal at node 16, say $I_4$. Note that the current generator is open as $V_{i+}=V_{i-}=0$.

From fig. 5.4 (b), it can be seen the output common-mode impedance of the OTA is

$$Z_{ocm} = \frac{v(\text{node 16})}{I_4} \quad (5.9)$$

The phase of $Z_{ocm}$ shows it is capacitive, so

$$Z_{ocm} = R_{ocm} \| \frac{1}{sC_{ocm}} \quad (5.10)$$


The extraction circuit structure is shown as Fig. 5.4 (a). Refer to Fig. 5.5. Let $V_{in3}=1\text{mV}$, $V_{in4}=-1\text{mV}$ and $RL1=RL2=350$ ohms. PSPICE was used to extract the current into terminal 16, say $I_5$. The current which passes through the output differential mode impedance, say $I_6$, is

$$I_6 = I_5 - \frac{v(\text{node 16})}{Z_{ocm}}$$

where $Z_{ocm}$ has been calculated in part D.
Therefore, the output differential-mode impedance of the OTA is

\[ Z_{\text{odm}} = \frac{v(\text{node } 16) - v(\text{node } 12)}{I_6 - \frac{v(\text{node } 16)}{Z_{\text{ocm}}}} \quad (5.11) \]

Again, it is found \( Z_{\text{odm}} \) is capacitive:

\[ Z_{\text{odm}} = R_{\text{odm}} // \frac{1}{sC_{\text{odm}}} \quad (5.12) \]

**F. Extraction of Output Impedance of the Specific OTA Circuit.**

The extraction circuit structure is shown as Fig. 5.4 (a). Let \( V_{\text{in3}} = 1 \text{mV}, V_{\text{in4}} = -1 \text{mV} \) and \( R_L = 350 \text{ ohms} \). PSPICE was used to extract the current into terminal 16, say \( I_7 \).

Then, the output impedance of the OTA is

\[ Z_{\text{odm}} = \frac{v(\text{node } 16) - v(\text{node } 12)}{I_7} \quad (5.13) \]

Table 5.1 (b) shows the extracted values of components of the proposed OTA model for \( g_{mT} = 2.48 \text{ millimhos} \) at frequency \( f_0 = 1.14 \text{ GHz} \) which is the center frequency of the bandpass filter on the transistor level. Fig. 5.6 and Fig 5.7 show the comparison of the input impedances and the output impedances of the OTA done with a simulation on the
transistor level with the results using the proposed model. The input impedances and the output impedances of the OTA transistor level simulation and the OTA model are very close. This means the extraction method is proper. Fig. 5.8 shows the input impedance of the OTA device is not affected by output loads $R_{L1}$ and $R_{L2}$. Therefore, no connecting elements from output to input are needed in the model. Note that $C_{ocm}$ and $R_{ocm}$ are negative because of the common mode amplifier which is connected to the output terminals inside the chip (refer to Table 5.1 (b)).
Fig. 5.1 A proposed OTA model which includes terminal impedances.
Equivalent circuit looking from the input port of the OTA

(a)

(b)

Fig. 5.2  
(a) Circuit for the extraction of the input impedance of the OTA and (b) The equivalent circuit of the input port.
Fig. 5.3 Extraction of the input differential mode impedance of the OTA.
Fig. 5.4  
(a) Circuit for finding the output impedance of an OTA.  
(b) The equivalent circuit at the output terminals.
Fig. 5.5 Extraction of the output differential mode impedance of the OTA.
Table 5.1 (a) The extracted values of components of the proposed OTA model for $g_mT = 2.48$ millimhos at $f_0=1.05$ GHz.

<table>
<thead>
<tr>
<th>Components</th>
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<td>Cicm</td>
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</tr>
<tr>
<td>Ricm</td>
<td>522600 ohms</td>
</tr>
<tr>
<td>Ridm</td>
<td>10340 ohms</td>
</tr>
<tr>
<td>Cidm</td>
<td>0.008633 pF</td>
</tr>
<tr>
<td>Cocm</td>
<td>-0.06458 pF</td>
</tr>
<tr>
<td>Rocm</td>
<td>-1731 ohms</td>
</tr>
<tr>
<td>Codm</td>
<td>0.07474 pF</td>
</tr>
<tr>
<td>Rodm</td>
<td>3290 ohms</td>
</tr>
<tr>
<td>$g_mT$</td>
<td>2.48 millimhos</td>
</tr>
</tbody>
</table>
### Table 5.1 (b) The extracted values of components of the proposed OTA model for $g_{mT} = 2.48$ millimhos at $f_0=1.14$ GHz.

<table>
<thead>
<tr>
<th>Components</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cicm</td>
<td>0.02725 pF</td>
</tr>
<tr>
<td>Ricm</td>
<td>1223000 ohms</td>
</tr>
<tr>
<td>Ridm</td>
<td>10310 ohms</td>
</tr>
<tr>
<td>Cidm</td>
<td>0.008613 pF</td>
</tr>
<tr>
<td>Cocm</td>
<td>-0.03004 pF</td>
</tr>
<tr>
<td>Rocm</td>
<td>-1854 ohms</td>
</tr>
<tr>
<td>Codm</td>
<td>0.05749 pF</td>
</tr>
<tr>
<td>Rodm</td>
<td>3512 ohms</td>
</tr>
<tr>
<td>$g_{mT}$</td>
<td>2.48 millimhos</td>
</tr>
</tbody>
</table>


<table>
<thead>
<tr>
<th>Components</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cicm</td>
<td>0.02727 pF</td>
</tr>
<tr>
<td>Rmem</td>
<td>-1487000 ohms</td>
</tr>
<tr>
<td>Ridm</td>
<td>10270 ohms</td>
</tr>
<tr>
<td>Cidm</td>
<td>0.008617 pF</td>
</tr>
<tr>
<td>Cocm</td>
<td>0.002144 pF</td>
</tr>
<tr>
<td>Rocm</td>
<td>-2053 ohms</td>
</tr>
<tr>
<td>Cprm</td>
<td>0.04146 pF</td>
</tr>
<tr>
<td>Rdom</td>
<td>3859 ohms</td>
</tr>
<tr>
<td>gmT</td>
<td>2.48 millimhos</td>
</tr>
</tbody>
</table>

Table 5.1 (c) The extracted values of components of the proposed OTA model for $g_{mT} = 2.48$ millimhos at $f_0 = 1.25$ GHz.
Fig. 5.6  (a) Magnitudes and (b) Phases of the input impedance of the OTA simulated on the transistor level and with the proposed model.
Fig. 5.7 (a) Magnitudes and (b) Phases of the output impedance of the OTA simulated on the transistor level and with the proposed model.
Fig. 5.8 (a) Magnitudes and (b) Phases of the input impedance of the OTA for different output resistors $R_{L1}$ and $R_{L2}$ simulated on the transistor level. Note: $R_{L1} = R_{L2}$. 
5.4 Simulations of the Bandpass Filter Utilizing the Proposed OTA Model

We use the proposed OTA model to get an equivalent circuit of the biquad $g_m$-RC bandpass active filter. The equivalent circuit is shown in Fig. 5.9. The PSPICE input circuit file is included in Appendix D. Fig. 5.10 shows the output voltage of the PSPICE simulation of this as the OTA output resistors $R_{L1}$, $R_{L2}$, and $R_{L3}$ are changed. The phases of the bandpass filter shown in Fig. 5.10 (b) are smooth and well behaved (zero degree and monotonically decreasing) around the center frequencies. Therefore, the filter is stable. Fig. 5.11 shows the best results of the equivalent circuit simulation which has quality factor $Q = 34$ and a center frequency $f_0 = 1.15$ GHz (after setting each OTA output load to $R_{L1}=R_{L2}=R_{L3}=350$ ohms). Fig. 5.12 shows the output voltage of the equivalent circuit when $R_1$ is varied. In Fig. 5.13, $C_1$ is varied, while in Fig. 5.14, the variable is $C_2$. These results are closer to the behavior of the actual OTA than the simpler model of Fig. 4.3 is.

From Table 5.1 (a) - (c), we see certain resistor and capacitor values change with frequency. All simulations above are done using the values with $f_0 = 1.14$ GHz. This causes a problem at other frequencies. This is one reason why the model doesn’t agree with a device level simulation completely.

In the next chapter, a modified biquad $g_m$-RC bandpass active filter is produced by taking account of the characteristics of the CMA in an actual OTA, which greatly improves the quality factor $Q$. 
Fig. 5.9  The equivalent circuit of the biquad $g_m$-RC bandpass active filter utilizing the proposed OTA model in Fig. 5.1.
Fig. 5.10 Effect of output terminal resistors $R_{L1}$, $R_{L2}$, and $R_{L3}$ on (a) the magnitude and (b) the phase of the output voltage gain of the biquad $g_m$-RC filter utilizing the proposed OTA model.

($g_{m1}=g_{m2}=g_{m3}=2.48$ millimhos, $C_1=1$ pF, $C_2=0.01$ pF, $R_1=60$ ohms)
Fig. 5.11  
(a) The magnitude and (b) the phase of output voltage gain of the biquad $g_m$-RC bandpass filter utilizing the proposed OTA model. 
($g_{mT1}=g_{mT2}=g_{mT3}=2.48$ millimhos, $C_1=1$ pF, $C_2=0.01$ pF, $R_1=60$ ohms, $R_{L1}=R_{L2}=R_{L3}=350$ ohms)
Fig. 5.12 Effect of resistor $R_1$ on (a) the magnitude and (b) the phase of the output voltage gain of the bandpass filter utilizing the proposed OTA model. ($g_{mT1}=g_{mT2}=g_{mT3}=2.48$ millimhos, $C_1=1$ pF, $C_2=0.01$ pF, $R_{L1}=R_{L2}=R_{L3}=350$ ohms)
Fig. 5.13  Effect of capacitor $C_1$ on (a) the magnitude and (b) the phase of the output voltage gain of the bandpass filter utilizing the proposed OTA model. ($g_{mT1}=g_{mT2}=g_{mT3}=2.48$ millimhos, $R_1=60$ ohms, $C_2=0.01$ pF, $R_{L1}=R_{L2}=R_{L3}=350$ ohms)
Fig. 5.14  Effect of capacitor $C_2$ on (a) the magnitude and (b) the phase of the output voltage of the bandpass filter utilizing the proposed OTA model. ($g_{m1}=g_{m2}=g_{m3}=2.48$ millimhos, $R_1=60$ ohms, $C_1=1$ pF, $R_{L1}=R_{L2}=R_{L3}=350$ ohms)
CHAPTER 6

A MODIFICATION OF THE $g_m$-RC BANDPASS FILTER TO OBTAIN HIGHER $Q$

In this chapter, we discuss the modification of the $g_m$-RC bandpass active filter in order to get a better $Q$ performance. The simulation results of the filter on a transistor level and the proposed OTA model are presented and contrasted. Finally, the limitation of the proposed OTA model is presented.

6.1 The Modified Biquad $g_m$-RC Bandpass Active Filter

In order to maintain a high quality factor $Q$ in the $g_m$-RC bandpass filter, we need to choose resistor $R_1$ to be less than $R_{L2}$. (refer to Fig. 4.2 and Fig. 4.5) However, the common mode feedback amplifier in the OTA tends to drive $v_{cmout} = (v_{16} + v_{12}) / 2$ to zero. (refer to Eq. (3.30)) If resistor $R_1$ is too small, it requires more current than $R_{L2}$ does to make $v_{16} = -v_{12}$. The transistor $Q_{16}$ in the OTA can go into saturation attempting to deliver this extra current. Note that all of the transistors in the filter need to be in the active region for the filter to work correctly. Therefore, there is a limitation in the quality factor $Q$ of the filter. One way to overcome the limitation on $Q$ is to connect the resistor $R_1$ between node 5 and node 7 instead of node 5 and ground. (see Fig. 6.1) In this way, we can choose a smaller resistor $R_1$ in order to get a wider range of $R_{L2}$ and yet not make $Q_{16}$ go into saturation.

The component values chosen for the best results for the filter are shown in Table 6.1. We chose $R_1 = 16$ ohms and $R_{L1} = R_{L2} = R_{L3} = 20$ ohms. Fig. 6.2 to Fig. 6.6 show
the results of PSPICE simulations on a transistor level. The quality factor $Q$ is up to 1141 and the center frequency $f_0$ is 1.14GHz. The phase plot given in Fig. 6.2 (b) shows that the modified bandpass filter is stable. Obviously, the quality factor $Q$ shows great improvement. (refer to Fig. 4.10)

Fig. 6.3 shows the output voltage of the filter as the resistor $R_1$ is changed. The center frequencies in each case are almost same. Fig. 6.4 shows the output voltage of the filter if $R_{L1} = R_{L2} = R_{L3} = R_L$. The curves were plotted for several values of $R_L$. The center frequencies for all curves are almost same here, too. Fig. 6.5 shows the output voltage of the filter as the capacitor $C_1$ is changed and Fig. 6.6 shows the output voltage of the filter using different values of $C_2$.

It is worthwhile mentioning that if one wants to use the filter, the capacitors $C_1$ and $C_2$ can be adjusted first to get the correct center frequency according to Eq. (4.9). Then adjusting the resistor $R_1$ and resistors $R_{L1}$, $R_{L2}$, and $R_{L3}$ gives the best quality factor $Q$ for this frequency. The filter can thus be optimized easily by choosing the components in this order.
Fig. 6.1 The schematic diagram of the proposed modified biquad \(g_m\)-RC bandpass filter.
<table>
<thead>
<tr>
<th>Devices</th>
<th>Parameter of Component</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resistor</td>
<td>RL1</td>
<td>20 ohms</td>
</tr>
<tr>
<td></td>
<td>RL2</td>
<td>20 ohms</td>
</tr>
<tr>
<td></td>
<td>RL3</td>
<td>20 ohms</td>
</tr>
<tr>
<td></td>
<td>R1</td>
<td>16 ohms</td>
</tr>
<tr>
<td>Capacitor</td>
<td>C1</td>
<td>1 PF</td>
</tr>
<tr>
<td></td>
<td>C2</td>
<td>0.01 PF</td>
</tr>
</tbody>
</table>

Table 6.1 The parameter values of components of the filter in Fig. 6.1.
(a) The magnitude and (b) the phase of the voltage gain of the filter in Fig. 6.1 simulated on the transistor level as $R_1$ is changed. ($R_{L1}=R_{L2}=R_{L3}=20$ ohms, $C_1=1\text{pF}$, $C_2=0.01\text{pF}$)
Fig. 6.4  Effect of output terminal resistors $R_{L1}$, $R_{L2}$, and $R_{L3}$ on (a) the magnitude and (b) the phase of voltage gain of the filter in Fig. 6.1 simulated on the transistor level. ($R_1=16$ ohms, $C_1=1pF$, $C_2=0.01pF$)
Fig. 6.5  (a) The magnitude and (b) the phase of the voltage gain of the filter in Fig. 6.1 simulated on the transistor level as $C_1$ is changed. ($R_{L1}=R_{L2}=R_{L3}=20 \text{ ohms}, R_1=16 \text{ ohms}, C_2=0.01\text{pF}$)
Fig. 6.6  (a) The magnitude and (b) the phase of the voltage gain of the filter in Fig. 6.1 simulated on the transistor level as $C_2$ is changed. ($R_{L1}=R_{L2}=R_{L3}=20$ ohms, $R_1=16$ ohms, $C_1=1$ pF)
6.2 PSPICE Simulations of the Modified Biquad $g_m$-RC Bandpass Active Filter Utilizing the Proposed OTA Model

The equivalent circuit of the modified filter utilizing the proposed OTA model is shown in Fig. 6.7. All values of components of the proposed OTA model have been shown in Table 5.1. The results of PSPICE simulations are shown in Fig. 6.8 through Fig. 6.12. Fig. 6.8 shows a quality factor $Q = 73$ and center frequency $f_0 = 1.15$GHz. Fig. 6.9 shows the output voltage of Fig. 6.7 as the resistor $R_1$ is changed. Fig. 6.10 shows the output voltage as the OTA output resistors $R_{L1}$, $R_{L2}$, and $R_{L3}$ are varied. Fig. 6.11 shows the output voltage of the circuit for different values of $C_1$, and Fig. 6.12 shows the output voltage of the equivalent circuit with capacitor $C_2$ as a parameter.
Fig. 6.7  The equivalent circuit of the modified biquad $g_m$-RC bandpass active filter utilizing the proposed OTA model.
Fig. 6.8  
(a) The magnitude and (b) the phase of the voltage gain of the equivalent circuit of Fig. 6.7. ($g_{m1}=g_{m2}=g_{m3}=2.48$ millimhos, $C_1=1$ pF, $C_2=0.01$ pF, $R_1=16$ ohms, $R_{L1}=R_{L2}=R_{L3}=20$ ohms)
Fig. 6.9  
(a) the magnitude and (b) the phase of the voltage gain of the equivalent circuit of Fig. 6.7 as $R_1$ is changed.  
($g_{m1}=g_{m2}=g_{m3}=2.48$ millimhos, $C_1=1$ pF, $C_2=0.01$ pF,  
$R_{L1}=R_{L2}=R_{L3}=20$ ohms)
Fig. 6.10  Effect of output terminal resistors $R_{L1}$, $R_{L2}$, and $R_{L3}$ on (a) the magnitude and (b) the phase of the voltage gain of the equivalent circuit of Fig. 6.7. ($g_{m1}=g_{m2}=g_{m3}=2.48$ millimhos, $C_1=1$ pF, $C_2=0.01$ pF, $R_1=16$ ohms)
Fig. 6.11 (a) the magnitude and (b) the phase of the voltage gain of the equivalent circuit of Fig. 6.7 as $C_1$ is changed.

($g_{mT1}=g_{mT2}=g_{mT3}=2.48$ millimhos, $R_1=16$ ohms, $C_2=0.01$ pF, $R_{L1}=R_{L2}=R_{L3}=20$ ohms)
(a) the magnitude and (b) the phase of the voltage gain of the equivalent circuit of Fig. 6.7 as $C_2$ is changed.

($g_mT_1 = g_mT_2 = g_mT_3 = 2.48$ millimhos, $R_1 = 16$ ohms, $C_1 = 1$ pF, $R_{L,1} = R_{L,2} = R_{L,3} = 20$ ohms)
6.3 The Strengths and Limitations of the Proposed OTA Model

From Figs. 4.4, 4.10 and 5.11, we see that the quality factor $Q$ of the filter utilizing the proposed OTA model is closer to the result simulated on the transistor level than when the ideal current source model is used for the OTA. Figs. 4.10, 5.11 and Figs. 6.2 and 6.8 shows the center frequency $f_0$ is 1.14 GHz simulated on the transistor level and 1.15 GHz with the proposed model. We see the proposed model predicts the $f_0$ of the filter simulated on the transistor level very well. According to Eq. (4.9), $C_1$ and $C_2$ affect the location of $f_0$. The larger $C_1$ and $C_2$ are, the smaller $f_0$ is. The model predicts the trend (see Figs. (4.12), (4.13), (5.13), (5.14) and Figs. (6.5), (6.6), (6.11), (6.12)). However, from Figs. 4.10, 5.11 and Figs. 6.2 and 6.8, we also realize the $Q$ of the filter calculated with the proposed model still has error. It appears that the reason is the variable capacitances of the common-mode and differential-mode output impedance. They are not constant, but strongly change with the frequency. Therefore, as the working frequency of the filter is changed, the extracted values of components of the proposed OTA model also need to be changed to get very accurate results.

In this chapter, we have introduced a way to improve the quality factor $Q$ of the filter. The PSPICE simulations have been done at both the device level and with the proposed model. The results shows the quality factor $Q$ of the filter is up to 1141 in the device level simulation and is 73 with the proposed model. Thus the proposed OTA model gives good results, but has the limitation that some of the model's parameters vary with frequency.
CHAPTER 7
CONCLUSION

7.1 Summary of the Thesis

We have used GaAs HBTs in a $g_m$ adjustable high-frequency fully differential OTA. The transconductance $g_{mT}$ of the OTA can be changed by setting some external voltages and a pair of resistors connected to the OTA. Simulations show the amplifier is broadband from DC to 42GHz, the maximum linear input range is $\pm 1.5$V and the transconductance coefficient $g_{mT}$ has value up to 2.48 millimhos. A built in feedback amplifier minimizes the common mode output voltages, and provides stabilization of the common mode signal levels of the OTA.

The amplifier was simulated for a biquad $g_m$-RC bandpass active filter application. The circuit analysis and the PSPICE simulation show the filter with the simple current generator model has a center frequency $f_0 = 3.98$ GHz and a quality factor $Q = 0.68$ (refer to Fig. 4.4). The filter was also simulated on the transistor level. The best result of simulations of the filter shows that the center frequency is around 1.14 GHz and the quality factor $Q$ is up to 139 (refer to Fig. 4.10). In addition, the $Q$ of the filter simulated on the transistor level can be controlled by adjusting the OTA output terminal resistors $R_L$. In order to get accurate design verification and to implement a procedure for accurate design of the filter, a proposed OTA model which includes nonidealities was developed. PSPICE simulations show the filter with the proposed model has the $f_0 = 1.15$ GHz and the $Q = 34$ (refer to Fig. 5.11). The results are closer to the more thorough simulation than the simple OTA model’s predictions.
By taking account of the characteristics of the actual OTA, a modified filter was produced. The PSPICE simulations on the transistor level show the Q of the filter is up to 1141 and \( f_0 = 1.15 \text{ GHz} \). The quality factor Q is improved greatly. The simulations with the proposed model show \( Q = 73 \) and \( f_0 = 1.15 \text{ GHz} \).

The proposed model was able to predict center frequency as well as the effect of the OTA's load resistors in the filter. However, there is the limitation that some of the model's parameters vary with frequency. This is the reason why the model can only estimate the Q, but doesn't agree with a device level simulation completely.

7.2 Suggestions for Future Study

In this thesis, we tested the OTA and obtained excellent results. It is worthwhile to do more investigation to apply the OTA in different kinds of filters such as lowpass, highpass, and bandreject active filters. In addition, the characteristics of the common mode feedback amplifier of the OTA which can be used in improving the quality factor Q of the bandpass filter is another interesting point requiring more study. This work can be applied not only to HBT devices but also the traditional silicon devices. This method of achieving high Q is a lot simpler than using the traditional very complex Q-control circuits presented in the literature[1, 25].
APPENDIX A

PSPICE CIRCUIT FILE FOR 42 GHz LINEAR OPERATIONAL TRANSCONDUCTANCE AMPLIFIER (OTA)

* 42 GHz LINEAR OTA
*
Q1 3 1 5 HBT1 1
R1 5 6 180
C1 5 6 50FF
Q3 6 6 8 HBT1 0.5
Q5 9 6 10 HBT1 0.5
Q7 12 11 9 HBT1 1
Q9 12 13 14 HBT2 0.5
R9 3 14 290
R19 3 15 290
Q10 16 13 15 HBT2 0.5
Q8 16 11 17 HBT1 1
Q6 17 18 10 HBT1 0.5
Q4 18 18 8 HBT1 0.5
R11 19 18 180
C11 19 18 50FF
Q2 3 2 19 HBT1 1
*
* COMMON-MODE FEEDBACK CIRCUIT
*
R15 3 69 800
Q15 62 62 69 HBT2 1
Q11 62 16 68 HBT1 1
R41 68 67 1.6K
R42 66 67 1.6K
Q13 13 0 66 HBT1 1
Q16 13 62 61 HBT2 1
R25 3 61 800
Q14 13 0 65 HBT1 1
R43 65 64 1.6K
R44 63 64 1.6K
Q12 62 12 63 HBT1 1
*
* POWER SUPPLY
*
VCC 3 0 5
VEE 4 0 -5
VB 0 11 0.5
VA1 34 0 3.8
VA2 44 0 4.4
VA3 54 0 1.5
VA4 74 0 1.5
*
* INPUT VOLTAGE SOURCE *
*
RS 105 0 1MEG
E1 1 0 105 0 0.5
E2 2 0 105 0 -0.5
VIN 105 0 DC 0 AC 1.0
*
* CURRENT SOURCE1 *
*
QE1 8 31 32 HBT1 0.5
QE2 31 32 4 HBT1 0.5
QD1 32 32 4 HBT1 2
QE3 31 0 33 HBT2 1
RE1 34 33 1K
*
* CURRENT SOURCE2 *
*
QE11 10 41 42 HBT1 0.5
QE12 41 42 4 HBT1 0.5
QD11 42 42 4 HBT1 2
QE13 41 0 43 HBT2 1
RE2 44 43 1K
*
* CURRENT SOURCE3 *
*
QE21 67 51 52 HBT1 0.5
QE22 51 52 4 HBT1 0.5
QD21 52 52 4 HBT1 2
QE23 51 0 53 HBT2 1
RE3 54 53 1K
*
* CURRENT SOURCE4 *
*
QE31 64 71 72 HBT1 0.5
QE32 71 72 4 HBT1 0.5
QD31 72 72 4 HBT1 2
QE33 71 0 73 HBT2 1
RE4 74 73 1K
*
* OUTPUT LOAD *
*
RL1 16 0 50
RL2 12 0 50
*
*
.MODEL HBT1 NPN(IS=1.93E-14 NF=1.05 VAF=20 ISE=1.44E-14 NE=1.17
+ISC=1.9E-13 NC=1.05 VAR=20 RB=15.6 RC=1.6 RE=1.34 CJ=0.096PF
+VJE=0.6 MJE=0.27 CJC=0.038PF FC=0.92 VJC=0.07 MJC=0.1 XCJC=0.3
+TF=1.5PS BF=1000 EG=1.43)
* .MODEL HBT2 PNP(IS=5.3E-18 NF=1.29 NE=1.7 ISE=1.6E-16 ISC=1.5E-10 +NC=1.29 RE=85 RB=0 RC=120 CJE=0.04PF VJC=0.102 VJE=1.147 MJE=0.411 +CJC=0.056PF FC=0.841 MJC=0.093 TF=3.7PS EG=1.43 BF=1000 VA=1000) *
* *.DC VIN -4.5 4.5 0.1 *.AC LIN 100 0.1G 100G *.OP *.PROBE *.END
APPENDIX B

PSPICE CIRCUIT FILE FOR TEST OF THE COMMON-MODE FEEDBACK CIRCUIT
OF THE OTA

* COMMON-MODE FEEDBACK CIRCUIT

R15 3 69 800
Q15 62 62 69 HBT2 1
Q11 62 160 68 HBT1 1
R41 68 67 1.6K
R42 66 67 1.6K
Q13 13 0 66 HBT1 1
Q16 13 62 61 HBT2 1
R25 3 61 800
Q14 13 0 65 HBT1 1
R43 65 64 1.6K
R44 63 64 1.6K
Q12 62 120 63 HBT1 1
Q9 12 13 14 HBT2 0.5
R9 3 14 290
R19 3 15 290
Q10 16 13 15 HBT2 0.5
RL1 12 0 50
RL2 16 0 50

* POWER SUPPLY

VCC 3 0 5
VEE 4 0 -5
VA3 54 0 2
VA4 74 0 2

* COMMON-MODE INPUT VOLTAGE SOURCE

RS 105 0 1MEG
E1 160 0 105 0 0.5
E2 120 0 105 0 0.5
VIN 105 0 DC 0 AC 1.0

* DIFFERENTIAL-MODE INPUT VOLTAGE SOURCE

*RS 105 0 1MEG
*E1 160 0 105 0 0.5
*E2 120 0 105 0 -0.5
*VIN 105 0 DC 0 AC 1.0
* CURRENT SOURCE3

*  
QE21 67 51 52  HBT1 0.5  
QE22 51 52 4  HBT1 0.5  
QD21 52 52 4  HBT1 2  
QE23 51 0 53  HBT2 1  
RE3 54 53 1K  
*  
* CURRENT SOURCE4  
*  
QE31 64 71 72  HBT1 0.5  
QE32 71 72 4  HBT1 0.5  
QD31 72 72 4  HBT1 2  
QE33 71 0 73  HBT2 1  
RE4 74 73 1K  
*  
* .MODEL HBT1 NPN(IS=1.93E-14 NF=1.05 VAF=20 ISE=1.44E-14 NE=1.17  
+ISC=1.9E-13 NC=1.05 VAR=20 RB=15.6 RC=1.6 RE=1.34 CJE=0.096PF  
+VJE=0.6 MJE=0.27 CJC=0.038PF FC=0.92 VJC=0.07 MJC=0.1 XCJC=0.3  
+TF=1.5PS BF=1000 EG=1.43)  
*  
*.MODEL HBT2 PNP(IS=5.3E-18 NF=1.29 NE=1.7 ISE=1.6E-16 ISC=1.5E-10  
+NC=1.29 RE=85 RB=0 RC=120 CJE=0.04PF VJC=0.102 VJE=1.147 MJE=0.411  
+CJC=0.056PF FC=0.841 MJC=0.093 TF=3.7PS EG=1.43 BF=1000 VA=1000)  
*  
*.DC VIN -4.5 4.5 0.1  
*.AC LIN 100 0.1G 100G  
*.OP  
*.PROBE  
*.END
APPENDIX C

PSPICE CIRCUIT FILE FOR THE SPECIFIC Biquad $g_m$-RC BANDPASS ACTIVE FILTER UTILIZING 3 OTA DEVICES

* THE SPECIFIC Biquad $g_m$-RC BANDPASS ACTIVE FILTER UTILIZING 3 OTA * DEVICES *

.SUBCKT OTAB 1 2 16 12
Q1 3 1 5 HBT1 1
R1 5 6 180
C1 5 6 50FF
Q3 6 6 8 HBT1 0.5
Q5 9 6 10 HBT1 0.5
Q7 12 11 9 HBT1 1
Q9 12 13 14 HBT2 0.5
R9 3 14 290
R19 3 15 290
Q10 16 13 15 HBT2 0.5
Q8 16 11 17 HBT1 1
Q6 17 18 10 HBT1 0.5
Q4 18 18 8 HBT1 0.5
R11 19 18 180
C11 19 18 50FF
Q2 3 2 19 HBT1 1

* COMMON-MODE FEEDBACK CIRCUIT *

R15 3 69 800
Q15 62 62 69 HBT2 1
Q11 62 16 68 HBT1 1
R41 68 67 1.6K
R42 66 67 1.6K
Q13 13 0 66 HBT1 1
Q16 13 62 61 HBT2 1
R25 3 61 800
Q14 13 0 65 HBT1 1
R43 65 64 1.6K
R44 63 64 1.6K
Q12 62 12 63 HBT1 1

* POWER SUPPLY *

VCC 3 0 5
VEE 4 0 -5
VB 0 11 0.5
VA1 34 0 3.8
\begin{verbatim}
VA2  44  0  4.4
VA3  54  0  1.5
VA4  74  0  1.5

* CURRENT SOURCE 1
*
QE1  8  31  32  HBT1 0.5
QE2  31  32  4  HBT1 0.5
QD1  32  32  4  HBT1 2
QE3  31  0  33  HBT2 1
RE1  34  33  1K

* CURRENT SOURCE 2
*
QE11 10  41  42  HBT1 0.5
QE12  41  42  4  HBT1 0.5
QD11  42  42  4  HBT1 2
QE13  41  0  43  HBT2 1
RE2  44  43  1K

* CURRENT SOURCE 3
*
QE21  67  51  52  HBT1 0.5
QE22  51  52  4  HBT1 0.5
QD21  52  52  4  HBT1 2
QE23  51  0  53  HBT2 1
RE3  54  53  1K

* CURRENT SOURCE 4
*
QE31  64  71  72  HBT1 0.5
QE32  71  72  4  HBT1 0.5
QD31  72  72  4  HBT1 2
QE33  71  0  73  HBT2 1
RE4  74  73  1K

.Model  HBT1 NPN(IS=1.93E-14 NF=1.05 VAF=20 ISE=1.44E-14 NE=1.17
+ISC=1.9E-13 NC=1.05 VAR=20 RB=15.6 RC=1.6 RE=1.34 CJE=0.096PF
+VJE=0.6 MJE=0.27 CJC=0.038PF FC=0.92 VJC=0.07 MJC=0.1 XCJC=0.3
+TF=1.5PS BF=1000 EG=1.43)

.Model  HBT2 PNP(IS=5.3E-18 NF=1.29 NE=1.7 ISE=1.6E-16 ISC=1.5E-10
+NC=1.29 RE=85 RB=0 RC=120 CJE=0.04PF VJC=0.102 VJE=1.147 MJE=0.411
+CJC=0.056PF FC=0.841 MJC=0.093 TF=3.7PS EG=1.43 BF=1000 VA=1000)

. ENDS
*

XB1  0  3  2  6  OTAB
XB2  3  0  5  7  OTAB
\end{verbatim}
XB3 2 5 3 8 OTAB
R2 6 0 350
R3 7 0 350
R4 8 0 350
C1 3 0 1PF
C2 3 0 0.01PF
R1 5 0 60
VIN 1 0 AC 1
.AC LIN 300 1.08G 1.2G
.OP
.PROBE
.END
APPENDIX D

PSPICE CIRCUIT FILE FOR THE EQUIVALENT CIRCUIT OF THE SPECIFIC
BIQUAD $g_m$-RC BANDPASS ACTIVE FILTER UTILIZING THE PROPOSED OTA
MODEL

* THE EQUIVALENT CIRCUIT OF THE SPECIFIC Biquad $g_m$-RC BANDPASS
* ACTIVE FILTER UTILIZING THE PROPOSED OTA MODEL
*

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.ENDS

* 

| XB1 | 0  | 3  | 2  | 6  | EQUC |
| XB2 | 3  | 0  | 5  | 7  | EQUC |
| XB3 | 2  | 5  | 3  | 8  | EQUC |
| R2  | 6  | 0  | 350 |
| R3  | 7  | 0  | 350 |
| R4  | 8  | 0  | 350 |
| C1  | 3  | 0  | 1PF |
| C2  | 3  | 0  | 0.01PF |
| R1  | 5  | 0  | 60  |
| VIN | 1  | 0  | AC  | 1  |

.AC LIN 500 1.0G 1.3G

.PROBE

.END
APPENDIX E
A FORTRAN PROGRAM TO CALCULATE THE INPUT COMMON-MODE IMPEDANCE OF THE OTA CIRCUIT FROM THE DATA EXTRACTED BY PSPICE

PROGRAM incmimp.f

C THIS IS A FORTRAN PROGRAM TO CALCULATE THE INPUT COMMON-MODE IMPEDANCE OF THE OTA CIRCUIT FROM THE DATA EXTRACTED BY PSPICE.
C
C AUTHOR : SUN SHAO-CHI

C
C COMPLEX V, C, Z, Y
REAL DATA1(30,5), VRI(30,3), IRI(30,3), ZRI(30,3), ZP(30,3), VA, IA
REAL R(30,3), YP(30,3)
OPEN (UNIT=2, FILE='incmimp.dat', STATUS='OLD')
C
C incmimp.dat INCLUDES VOLTAGE AND CURRENT OF INPUT NODE 1 OF THE OTA, WHICH EXTRACTED BY PSPICE.
C
C REWIND 2
DO 10 I=1, 30, 1
READ(2,100) (DATA(I,J),J=1,5)
10 CONTINUE
CLOSE(2)
100 FORMAT (E9.3,4E12.3)
C
C DATA(I,1) IS FREQUENCY.
C DATA(I,2) IS MAGNITUDE OF VOLTAGE OF INPUT NODE 1.
C DATA(I,3) IS PHASE OF VOLTAGE OF INPUT NODE 1.
C DATA(I,4) IS MAGNITUDE OF CURRENT OF INPUT NODE 1.
C DATA(I,5) IS PHASE OF CURRENT OF INPUT NODE 1.
C
DO 50 I=1,30,1
VRI(I,1)=DATA(I,1)
IRI(I,1)=DATA(I,1)
VA=DATA(I,3) * 3.14159 / 180
IA=DATA(I,5) * 3.14159 / 180
VRI(I,2)=DATA(I,2) * COS(VA)
VRI(I,3)=DATA(I,2) * SIN(VA)
IRI(I,2)=DATA(I,4) * COS(IA)
IRI(I,3)=DATA(I,4) * SIN(IA)
V=CMPLX(VRI(I,2), VRI(I,3))
C=CMPLX(IRI(I,2), IRI(I,3))
Z=V / C
Y=C / V
ZP(I,1)=DATA(I,1)
ZP(I,2)=REAL(Z)
ZP(I,3)=AIMAG(Z)
YP(I,1)=DATA(I,1)
YP(I,2)=REAL(Y)
YP(I,3)=AIMAG(Y)
ZRI(I,1)=DATA(I,1)
ZRI(I,2)=ABS(Z)
ZRI(I,3)=ATAN2(ZP(I,3), ZP(I,2)) * 180 / 3.14159
R(I,1)=DATA(I,1)
R(I,2)=1 / YP(I,2)
R(I,3)=YP(I,3) / (6.28318 * DATA(I,1))
50 CONTINUE
OPEN (UNIT=10, FILE='incmimp1.dat')
C
incmimp1.dat INCLUDES THE INPUT COMMON-MODE IMPEDANCE OF
C THE OTA.
C
REWIND 10
WRITE (10, 200)
200 FORMAT ('FREQ', 10X, 'REAL', 10X, 'IMAG')
DO 110 I=1, 30, 1
WRITE (10, 300) (ZP(I,J), J=1,3)
110 CONTINUE
CONTINUE
WRITE (10, 400)
WRITE (10, 400)
400 FORMAT ('FREQ', 10X, 'MAG', 10X, 'PHASE')
DO 500 I=1,30,1
WRITE (10,300) (ZRI(I,J), J=1,3)
500 CONTINUE
CONTINUE
WRITE(10,600)
WRITE(10,600)
600 FORMAT ('FREQ', 9X, 'RESISTANCE', 3X, 'CAPACITANCE')
DO 700 I=1,30,1
WRITE (10, 300) (R(I,J), J=1,3)
700 CONTINUE
CLOSE(10)
300 FORMAT (E10.4, 2E13.4)
END
APPENDIX  F

A FORTRAN PROGRAM TO CALCULATE THE INPUT DIFFERENTIAL MODE IMPEDANCE OF THE OTA CIRCUIT FROM THE DATA EXTRACTED BY PSPICE

```
C PROGRAM outdmimp.f
C THIS IS A FORTRAN PROGRAM TO CALCULATE THE INPUT DIFFERENTIAL MODE IMPEDANCE OF THE OTA CIRCUIT FROM THE DATA EXTRACTED BY PSPICE.
C AUTHOR: SUN SHAO-CHI
C
COMPLEX  V, C, ZCMC, ZDMC, C1, V2, V1, YDMC
REAL    DATA1(30,5), VRI(30,3), IRI(30,3), ZRI(30,3), ZP(30,3), VA, IA
REAL    R(30,3), ZCM(30,3), VR1(30,3), IR1(30,3), DATA2(30,5), VV, II
REAL    YP(30,3)
OPEN (UNIT=2, FILE='indmimpl.dat', STATUS='OLD')
C
C indmimpl.dat INCLUDES THE VOLTAGE AND CURRENT OF INPUT NODE 1 OF THE OTA, WHICH EXTRACTED BY PSPICE.
C
REWIND 2
DO 10 I=1, 30, 1
READ (2,100) (DATA1(I,J),J=1,5)
C
DATA1(I,1) IS FREQUENCY.
DATA1(I,2) IS MAGNITUDE OF VOLTAGE OF INPUT NODE 1.
DATA1(I,3) IS PHASE OF VOLTAGE OF INPUT NODE 1.
DATA1(I,4) IS MAGNITUDE OF CURRENT OF INPUT NODE 1.
DATA1(I,5) IS PHASE OF CURRENT OF INPUT NODE 1.
10 CONTINUE
CLOSE(2)
OPEN (UNIT=3, FILE='indmimp12.dat', STATUS='OLD')
C
C indmimp12.dat INCLUDES THE VOLTAGE AND CURRENT OF INPUT NODE 2 OF THE OTA, WHICH EXTRACTED BY PSPICE.
C
REWIND 3
DO 12 I=1, 30, 1
READ (3,100) (DATA2(I,J),J=1,5)
C
DATA2(I,1) IS FREQUENCY.
DATA2(I,2) IS MAGNITUDE OF VOLTAGE OF INPUT NODE 2.
DATA2(I,3) IS PHASE OF VOLTAGE OF INPUT NODE 2.
DATA2(I,4) IS MAGNITUDE OF CURRENT OF INPUT NODE 2.
DATA2(I,5) IS PHASE OF CURRENT OF INPUT NODE 2.
C```

12 CONTINUE
CLOSE(3)
100 FORMAT (E9.3,4E12.3)
OPEN (UNIT=20, FILE='incmimp2.dat', STATUS='OLD')
C
C incmimp2.dat INCLUDES THE INPUT COMMON-MODE IMPEDANCE OF
C THE OTA.
C
REWIND 20
DO 11 I=1, 30, 1
READ(20,300) (ZCM(I,J),J=I,3)
C
C ZCM(I,1) IS FREQUENCY.
C ZCM(I,2) IS REAL PART OF THE INPUT COMMON-MODE IMPEDANCE.
C ZCM(I,3) IS IMAGINARY PART OF THE INPUT COMMON-MODE
C IMPEDANCE.
C
11 CONTINUE
CLOSE(20)
D050
DO 50 I=1,30,1
VRI(I,1)=DATA1(I,1)
IRI(I,1)=DATA1(I,1)
VA=DATA1(I,3) * 3.14159 / 180
IA=DATA1(I,5) * 3.14159 / 180
VRI(I,2)=DATA1(I,2) * COS(VA)
VRI(I,3)=DATA1(I,2) * SIN(VA)
IRI(I,2)=DATA1(I,4) * COS(IA)
IRI(I,3)=DATA1(I,4) * SIN(IA)
V=CMPLX(VRI(I,2), VRI(I,3))
C=CMPLX(IRI(I,2), IRI(I,3))
VRI1(I,1)=DATA2(I,1)
IRI1(I,1)=DATA2(I,1)
VV=DATA2(I,3) * 3.14159 / 180
II=DATA2(I,5) * 3.14159 / 180
VRI1(I,2)=DATA2(I,2) * COS(VV)
VRI1(I,3)=DATA2(I,2) * SIN(VV)
IRI1(I,2)=DATA2(I,4) * COS(II)
IRI1(I,3)=DATA2(I,4) * SIN(II)
V1=CMPLX(VRI1(I,2), VRI1(I,3))
C1=CMPLX(IRI1(I,2), IRI1(I,3))
V2=V-V1
ZMC=CMPLX(ZCM(I,2), ZCM(I,3))
ZDMC=(V2 * ZMC) / (C * ZMC - V)
YDMC=1 / ZDMC
ZP(I,1)=DATA1(I,1)
ZP(I,2)=REAL(ZDMC)
ZP(I,3)=AIMAG(ZDMC)
YP(I,1)=DATA1(I,1)
YP(I,2)=REAL(YDMC)
YP(I,3)=AIMAG(YDMC)
ZRI(I,1)=DATA1(I,1)
ZRI(I,2)=ABS(ZDMC)
ZRI(I,3)=ATAN2(ZP(I,3), ZP(I,2)) * 180 / 3.14159
R(I,1)=DATA1(I,1)
R(I,2)=1 / YP(I,2)
R(I,3)=YP(I,3) / (6.28318 * DATA1(I,1))
50 CONTINUE
OPEN (UNIT=10, FILE='indmimp1.dat')
C
C      INDMIMP1.DAT INCLUDES THE INPUT DIFFERENTIAL-MODE IMPEDANCE
C      OF THE OTA.
C
REWRITE 10
WRITE (10, 200)
200 FORMAT ('FREQ', 10X, 'REAL', 10X, 'IMAG')
DO 110 I=1, 30, 1
   WRITE (10, 300) (ZP(I,J), J=1,3)
110 CONTINUE
WRITE (10, 400)
400 FORMAT ('FREQ', 10X, 'MAG', 10X, 'PHASE')
DO 500 I=1, 30, 1
   WRITE (10, 300) (ZRI(I,J), J=1,3)
500 CONTINUE
WRITE(10, 600)
600 FORMAT ('FREQ', 9X, 'RESISTANCE', 3X, 'CAPACITANCE')
DO 700 I=1, 30, 1
   WRITE (10, 300) (R(I,J), J=1,3)
700 CONTINUE
CLOSE(10)
300 FORMAT (E10.4, 2E13.4)
END
APPENDIX G

A FORTRAN PROGRAM TO CALCULATE THE INPUT IMPEDANCE OF THE OTA CIRCUIT FROM THE DATA EXTRACTED BY PSPICE

PROGRAM inimp.f
THIS IS A FORTRAN PROGRAM TO CALCULATE THE INPUT IMPEDANCE OF THE OTA CIRCUIT FROM THE DATA EXTRACTED BY PSPICE.
AUTHOR: SUN SHAO-CHI

COMPLEX V, C, Z, C1, V2, V1, Y
REAL DATA1(30,5), VRI(30,3), IRI(30,3), ZRI(30,3), ZP(30,3), VA, IA
REAL R(30,3), VRI1(30,3), IRI1(30,3), DATA2(30,5), VV, II
REAL YP(30,3)
OPEN (UNIT=2, FILE='indmimp13.dat', STATUS='OLD')

REWIND 2
DO 10 I=1, 30, 1
READ (2,100) (DATA1(I,J),J=1,5)

DATA1(I,1) IS FREQUENCY.
DATA1(I,2) IS MAGNITUDE OF VOLTAGE OF INPUT NODE 1.
DATA1(I,3) IS PHASE OF VOLTAGE OF INPUT NODE 1.
DATA1(I,4) IS MAGNITUDE OF CURRENT OF INPUT NODE 1.
DATA1(I,5) IS PHASE OF CURRENT OF INPUT NODE 1.
CONTINUE
CLOSE(2)
OPEN (UNIT=3, FILE='indmimp14.dat', STATUS='OLD')

REWIND 3
DO 12 I=1, 30, 1
READ (3,100) (DATA2(I,J),J=1,5)

DATA2(I,1) IS FREQUENCY.
DATA2(I,2) IS MAGNITUDE OF VOLTAGE OF INPUT NODE 2.
DATA2(I,3) IS PHASE OF VOLTAGE OF INPUT NODE 2.
DATA2(I,4) IS MAGNITUDE OF CURRENT OF INPUT NODE 2.
DATA2(I,5) IS PHASE OF CURRENT OF INPUT NODE 2.
CONTINUE
CLOSE(3)

FORMAT (E9.3,4E12.3)
DO 50  I=1,30,1
VRI(I,1)=DATA1(I,1)
IRI(I,1)=DATA1(I,1)
VA=DATA1(I,3) * 3.14159 / 180
IA=DATA1(I,5) * 3.14159 / 180
VRI(I,2)=DATA1(I,2) * COS(VA)
VRI(I,3)=DATA1(I,2) * SIN(VA)
IRI(I,2)=DATA1(I,4) * COS(IA)
IRI(I,3)=DATA1(I,4) * SIN(IA)
V=CMPLX(VRI(I,2), VRI(I,3))
C=CMPLX(IRI(I,2), IRI(I,3))
VRII(I,1)=DATA2(I,1)
IRI1(I,1)=DATA2(I,1)
VV=DATA2(I,3) * 3.14159 / 180
II=DATA2(I,5) * 3.14159 / 180
VRII(I,2)=DATA2(I,2) * COS(VV)
VRII(I,3)=DATA2(I,2) * SIN(VV)
IRI1(I,2)=DATA2(I,4) * COS(II)
IRI1(I,3)=DATA2(I,4) * SIN(II)
V1=CMPLX(VRII(I,2), VRII(I,3))
C1=CMPLX(IRI1(I,2), IRI1(I,3))
V2=V-V1
Z=V2 / C
Y=1 / Z
ZP(I,1)=DATA1(I,1)
ZP(I,2)=REAL(Z)
ZP(I,3)=AIMAG(Z)
YP(I,1)=DATA1(I,1)
YP(I,2)=REAL(Y)
YP(I,3)=AIMAG(Y)
ZRI(I,1)=DATA1(I,1)
ZRI(I,2)=ABS(Z)
ZRI(I,3)=ATAN2(ZP(I,3), ZP(I,2)) * 180 / 3.14159
R(I,1)=DATA1(I,1)
R(I,2)=1 / YP(I,2)
R(I,3)=YP(I,3) / (6.28318 * DATA1(I,1))
CONTINUE
OPEN (UNIT=10, FILE='inimp2.dat')
inimp2.dat INCLUDES THE INPUT IMPEDANCE OF THE OTA.
REWIND 10
WRITE (10, 200)
FORMAT ('FREQ', 10X, 'REAL', 10X, 'IMAG')
DO 110  I=1,30,1
WRITE (10, 300) (ZP(I,J), J=1,3)
CONTINUE
WRITE (10, 400)
DO 500 I=1,30,1
WRITE (10,300) (ZRI(I,J), J=1,3)
500 CONTINUE
WRITE(10,600)
600 FORMAT ('FREQ', 9X, 'RESISTANCE', 3X, 'CAPACITANCE')
DO 700 I=1,30,1
WRITE (10,300) (R(I,J), J=1,3)
700 CONTINUE
CLOSE(10)
300 FORMAT (E10.4, 2E13.4)
END
APPENDIX H

A FORTRAN PROGRAM TO CALCULATE THE OUTPUT COMMON MODE IMPEDANCE OF THE OTA CIRCUIT FROM THE DATA EXTRACTED BY PSPICE

C PROGRAM outcmimp.f
C THIS IS A FORTRAN PROGRAM TO CALCULATE THE OUTPUT COMMON-MODE IMPEDANCE OF THE OTA CIRCUIT FROM THE DATA EXTRACTED BY PSPICE.
C AUTHOR: SUN SHAO-CHI

COMPLEX V, C, Z, Y
REAL DATA(30,5), VRI(30,3), IRI(30,3), ZRI(30,3), ZP(30,3), VA, IA
REAL R(30,3), YP(30,3)
OPEN (UNIT=2, FILE='outcmimp.dat', STATUS='OLD')
outcmimp.dat INCLUDES VOLTAGE AND CURRENT OF OUTPUT NODE 16 OF THE OTA, WHICH EXTRACTED BY PSPICE.
REWIND 2
DO 10 I=1, 30, 1
READ(2,100) (DATA(I,J),J=1,5)
DATA(I,1) IS FREQUENCY.
DATA(I,2) IS MAGNITUDE OF VOLTAGE OF OUTPUT NODE 16.
DATA(I,3) IS PHASE OF VOLTAGE OF OUTPUT NODE 16.
DATA(I,4) IS MAGNITUDE OF CURRENT OF OUTPUT NODE 16.
DATA(I,5) IS PHASE OF CURRENT OF OUTPUT NODE 16.
CONTINUE
CLOSE(2)

10 FORMAT (E9.3,4(E12.3)
DO 50 I=1,30,1
VRI(I,1)=DATA(I,1)
IRI(I,1)=DATA(I,1)
VA=DATA(I,3) * 3.14159 / 180
IA=DATA(I,5) * 3.14159 / 180
VRI(I,2)=DATA(I,2) * COS(VA)
IRI(I,2)=DATA(I,2) * SIN(VA)
IRI(I,3)=DATA(I,3) * COS(IA)
IRI(I,3)=DATA(I,3) * SIN(IA)
V=CMPLX(VRI(I,2), VRI(I,3))
C=CMPLX(IRI(I,2), IRI(I,3))
Z=V / C
Y=C / V
ZP(I,1)=DATA(I,1)
ZP(I,2)=REAL(Z)
ZP(I,3) = AIMAG(Z)
YP(I,1) = DATA(I,1)
YP(I,2) = REAL(Y)
YP(I,3) = AIMAG(Y)
ZRI(I,1) = DATA(I,1)
ZRI(I,2) = ABS(Z)
ZRI(I,3) = ATAN2(ZP(I,3), ZP(I,2)) * 180 / 3.14159
R(I,1) = DATA(I,1)
R(I,2) = 1 / YP(I,2)
R(I,3) = YP(I,3) / (6.28318 * DATA(I,1))
50 CONTINUE
OPEN (UNIT=10, FILE='outcmimp1.dat')
C outcmimp1.dat INCLUDES THE OUTPUT COMMON-MODE IMPEDANCE OF
C THE OTA.
C
REWIND 10
WRITE (10, 200)
200 FORMAT ('FREQ', 10X, 'REAL', 10X, 'IMAG')
DO 110 I=1, 30, 1
WRITE (10, 300) (ZP(I,J), J=1,3)
110 CONTINUE
WRITE (10, 400)
400 FORMAT ('FREQ', 10X, 'MAG', 10X, 'PHASE')
DO 500 I=1,30,1
WRITE (10,300) (ZRI(I,J), J=1,3)
500 CONTINUE
WRITE (10, 600)
600 FORMAT ('FREQ', 9X, 'RESISTANCE', 3X, 'CAPACITANCE')
DO 700 I=1,30,1
WRITE (10, 300) (R(I,J), J=1,3)
700 CONTINUE
CLOSE(10)
300 FORMAT (E10.4, 2E13.4) END
APPENDIX I

A FORTRAN PROGRAM TO CALCULATE THE OUTPUT DIFFERENTIAL-MODE IMPEDANCE OF THE OTA CIRCUIT FROM THE DATA EXTRACTED BY PSPICE

PROGRAM outdmimp.f
THIS IS A FORTRAN PROGRAM TO CALCULATE THE OUTPUT DIFFERENTIAL MODE IMPEDANCE OF THE OTA CIRCUIT FROM THE DATA EXTRACTION BY PSPICE.
AUTHOR: SUN SHAO-CHI

COMPLEX V, C, ZCMC, ZDMC, C1, V2, V1, YDMC
REAL DATA1(30,5), VRI(30,3), IRI(30,3), ZRI(30,3), ZP(30,3), VA, IA
REAL R(30,3), ZCM(30,3), VR11(30,3), IR11(30,3), DATA2(30,5), VV, II
REAL YP(30,3)
OPEN (UNIT=2, FILE='outdmimp11.dat', STATUS='OLD')

outdmimp11.dat INCLUDES THE VOLTAGE AND CURRENT OF OUTPUT NODE 16 OF THE OTA, WHICH EXTRACTED BY PSPICE.

REWIND 2
DO 10 I=1, 30, 1
READ (2,100) (DATA1(I,J),J=1,5)

DATA1(I,1) IS FREQUENCY.
DATA1(I,2) IS MAGNITUDE OF VOLTAGE OF OUTPUT NODE 16.
DATA1(I,3) IS PHASE OF VOLTAGE OF OUTPUT NODE 16.
DATA1(I,4) IS MAGNITUDE OF CURRENT OF OUTPUT NODE 16.
DATA1(I,5) IS PHASE OF CURRENT OF OUTPUT NODE 16.

10 CONTINUE
CLOSE(2)
OPEN (UNIT=3, FILE='outdmimp12.dat', STATUS='OLD')

outdmimp12.dat INCLUDES THE VOLTAGE AND CURRENT OF OUTPUT NODE 12 OF THE OTA, WHICH EXTRACTED BY PSPICE.

REWIND 3
DO 12 I=1, 30, 1
READ (3,100) (DATA2(I,J),J=1,5)

DATA2(I,1) IS FREQUENCY.
DATA2(I,2) IS MAGNITUDE OF VOLTAGE OF OUTPUT NODE 12.
DATA2(I,3) IS PHASE OF VOLTAGE OF OUTPUT NODE 12.
DATA2(I,4) IS MAGNITUDE OF CURRENT OF OUTPUT NODE 12.
DATA2(I,5) IS PHASE OF CURRENT OF OUTPUT NODE 12.
CONTINUE
CLOSE(3)

OPEN (UNIT=20, FILE='outcmimp2.dat', STATUS='OLD')

outcmimp2.dat INCLUDES THE OUTPUT COMMON-MODE IMPEDANCE OF
THE OTA.

REWIND 20
DO 11 I=1, 30, 1
READ(20,300) (ZCM(I,J),J=1,3)

ZCM(I,1) IS FREQUENCY.
ZCM(I,2) IS REAL PART OF THE OUTPUT COMMON-MODE IMPEDANCE.
ZCM(I,3) IS IMAGINARY PART OF THE OUTPUT COMMON-MODE
IMPEDANCE.

CONTINUE
CLOSE(20)
DO 50 I=1,30,1
VRl(I,1)=DATA1(I,1)
IRI(I,1)=DATA1(I,1)
VA=DATA1(I,3) * 3.14159 / 180
IA=DATA1(I,5) * 3.14159 / 180
VRl(I,2)=DATA1(I,2) * COS(VA)
VRl(I,3)=DATA1(I,2) * SIN(VA)
IRI(I,2)=DATA1(I,4) * COS(IA)
IRI(I,3)=DATA1(I,4) * SIN(IA)
V=CMPLX(VRl(I,2), VRl(I,3))
C=CMPLX(IRI(I,2), IRI(I,3))
VRl1(I,1)=DATA2(I,1)
IRI1(I,1)=DATA2(I,1)
VV=DATA2(I,3) * 3.14159 / 180
II=DATA2(I,5) * 3.14159 / 180
VRl1(I,2)=DATA2(I,2) * COS(VV)
VRl1(I,3)=DATA2(I,2) * SIN(VV)
IRI1(I,2)=DATA2(I,4) * COS(II)
IRI1(I,3)=DATA2(I,4) * SIN(II)
V1=CMPLX(VRl1(I,2), VRl1(I,3))
C1=CMPLX(IRI1(I,2), IRI1(I,3))
V2=V-V1
ZCMC=CMPLX(ZCM(I,2), ZCM(I,3))
ZDMC=(V2 * ZCMC) / (C * ZCMC - V)
YDMC=1 / ZDMC
ZP(I,1)=DATA1(I,1)
ZP(I,2)=REAL(ZDMC)
ZP(I,3)=AIMAG(ZDMC)
YP(I,1)=DATA1(I,1)
YP(I,2)=REAL(YDMC)
YP(I,3)=AIMAG(YDMC)
ZRI(I,1)=DATA1(I,1)
ZRI(I,2)=ABS(ZDMC)
ZRI(I,3)=ATAN2(ZP(I,3), ZP(I,2)) * 180 / 3.14159
R(I,1)=DATA1(I,1)
R(I,2)=1 / YP(I,2)
R(I,3)=YP(I,3) / (6.28318 * DATA1(I,1))
50 CONTINUE
OPEN (UNIT=10, FILE='outdimp1.dat')
C outdimp1.dat INCLUDES THE OUTPUT DIFFERENTIAL-MODE IMPEDANCE OF THE OTA.
C
REWIN 10
WRITE (10, 200)
200 FORMAT ('FREQ', 10X, 'REAL', 10X, 'IMAG')
DO 110 I=1, 30, 1
WRITE (10, 300) (ZP(I,J), J=1,3)
110 CONTINUE
WRITE (10, 400)
400 FORMAT ('FREQ', 10X, 'MAG', 10X, 'PHASE')
DO 500 I=1,30,1
WRITE (10, 300) (ZRI(I,J), J=1,3)
500 CONTINUE
WRITE (10, 600)
600 FORMAT ('FREQ', 9X, 'RESISTANCE', 3X, 'CAPACITANCE')
DO 700 I=1,30,1
WRITE (10, 300) (R(I,J), J=1,3)
700 CONTINUE
CLOSE(10)
300 FORMAT (E10.4, 2E13.4)
END
APPENDIX  J

A FORTRAN PROGRAM TO CALCULATE THE OUTPUT IMPEDANCE OF THE
OTA CIRCUIT FROM THE DATA EXTRACTED BY PSPICE

PROGRAM outimp.f
C THIS IS A FORTRAN PROGRAM TO CALCULATE THE OUTPUT
C IMPEDANCE OF THE OTA CIRCUIT FROM THE DATA EXTRACTED BY
C PSPICE.
C AUTHOR: SUN SHAO-CHI
C
COMPLEX  V, C, Z, C1, V2, V1, Y
REAL   DATA1(30,5), VR(30,3), IR(30,3), ZP(30,3), VA, IA
REAL   R(30,3), VR1(30,3), IR1(30,3), DATA2(30,5), VV, II
REAL   YP(30,3)
OPEN (UNIT=2, FILE='outdmimp13.dat', STATUS='OLD')
C
C outdmimp13.dat INCLUDES THE VOLTAGE AND CURRENT OF OUTPUT
C NODE 16 OF THE OTA, WHICH EXTRACTED BY PSPICE.
C
REWIND 2
DO 10  I=1, 30, 1
READ (2,100) (DATA1(I,J),J=1,5)
C
C DATA1(I,1) IS FREQUENCY.
C DATA1(I,2) IS MAGNITUDE OF VOLTAGE OF OUTPUT NODE 16.
C DATA1(I,3) IS PHASE OF VOLTAGE OF OUTPUT NODE 16.
C DATA1(I,4) IS MAGNITUDE OF CURRENT OF OUTPUT NODE 16.
C DATA1(I,5) IS PHASE OF CURRENT OF OUTPUT NODE 16.
C
10 CONTINUE
CLOSE(2)
OPEN (UNIT=3, FILE='outdmimp14.dat', STATUS='OLD')
C
C outdmimp14.dat INCLUDES THE VOLTAGE AND CURRENT OF OUTPUT
C NODE 12 OF THE OTA, WHICH EXTRACTED BY PSPICE.
C
REWIND 3
DO 12  I=1, 30, 1
READ (3,100) (DATA2(I,J),J=1,5)
C
C DATA2(I,1) IS FREQUENCY.
C DATA2(I,2) IS MAGNITUDE OF VOLTAGE OF OUTPUT NODE 12.
C DATA2(I,3) IS PHASE OF VOLTAGE OF OUTPUT NODE 12.
C DATA2(I,4) IS MAGNITUDE OF CURRENT OF OUTPUT NODE 12.
C DATA2(I,5) IS PHASE OF CURRENT OF OUTPUT NODE 12.
C
CONTINUE
CLOSE(3)

FORMAT (E9.3,4E12.3)
DO 50 I=1,30,1
VRI(I,1)=DATA1(I,1)
IRI(I,1)=DATA1(I,1)
VA=DATA1(I,3) * 3.14159 / 180
IA=DATA1(I,5) * 3.14159 / 180
VRI(I,2)=DATA1(I,2) * COS(VA)
VRI(I,3)=DATA1(I,2) * SIN(VA)
IRI(I,2)=DATA1(I,4) * COS(IA)
IRI(I,3)=DATA1(I,4) * SIN(IA)
V=CMPLX(VRI(I,2), VRI(I,3))
C=CMPLX(IRI(I,2), IRI(I,3))
VRI(I,1)=DATA2(I,1)
IRI(I,1)=DATA2(I,1)
VV=DATA2(I,3) * 3.14159 / 180
II=DATA2(I,5) * 3.14159 / 180
VRI(I,2)=DATA2(I,2) * COS(VV)
VRI(I,3)=DATA2(I,2) * SIN(VV)
IRI(I,2)=DATA2(I,4) * COS(II)
IRI(I,3)=DATA2(I,4) * SIN(II)
V1=CMPLX(VRI(I,2), VRI(I,3))
C1=CMPLX(IRI(I,2), IRI(I,3))
V2=V-V1
Z=V2 / C
Y=1 / Z
ZP(I,1)=DATA1(I,1)
ZP(I,2)=REAL(Z)
ZP(I,3)=AIMAG(Z)
YP(I,1)=DATA1(I,1)
YP(I,2)=REAL(Y)
YP(I,3)=AIMAG(Y)
ZRI(I,1)=DATA1(I,1)
ZRI(I,2)=ABS(Z)
ZRI(I,3)=ATAN2(ZP(I,3), ZP(I,2)) * 180 / 3.14159
R(I,1)=DATA1(I,1)
R(I,2)=1 / YP(I,2)
R(I,3)=YP(I,3) / (6.28318 * DATA1(I,1))
CONTINUE
OPEN (UNIT=10, FILE='outimp2.dat')

outimp2.dat INCLUDES THE OUTPUT IMPEDANCE OF THE OTA.

REWIND 10
WRITE (10, 200)

400  FORMAT (‘FREQ’, 10X, ‘MAG’, 10X, ‘PHASE’)
    DO 500 I=1,30,1
    WRITE (10,300) (ZRI(I,J), J=1,3)
500  CONTINUE
    WRITE(10, 600)
    DO 700 I=1,30,1
    WRITE (10, 300) (R(I,J), J=1,3)
700  CONTINUE
    CLOSE(10)
300  FORMAT (E10.4, 2E13.4)
END
REFERENCES


[35] LIBRA, EEsof Corp., 5601 Lindero Canyon Rd., Westlake Village, CA 91362.


