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# Linear Voltage Controlled Oscillator Implementation in Electronically Variable Immittances

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**Abstract.** New realization schemes of electronically tunable inductor (L) and Frequency-Dependent Negative Resistance (D) type immittances using a *Current Feedback Amplifier* (CFA) and *Multiplication Mode Current Conveyor* (MMCC) composite *active building block* (ABB) are proposed. Applications of the immittances to the design of selective filters and LC-type *linear voltage controlled quadrature oscillator* (LVCQO) are presented. Experimental results based on PSPICE simulation and hardware design for a linear range of oscillation frequency ($f_o \sim 13.6$MHz) with satisfactory phase-noise figure on the oscillator wave response had been verified. Effects of the *Active Building Block* (ABB)–nodal imperfections are analyzed to be insignificant. The new ideas in this article are two types of immittance functions realizable in the same topology; appropriate frequency-domain selective responses are also presented with experimentation results.

**Key-words:** *Current Feedback Amplifier* (CFA); linear voltage control oscillator; quadrature; selective filter; synthetic inductor/FDNR.

## 1. Introduction

Design implementation of the inductor (L) and *Frequency Dependent Negative Resistance* (FDNR) (D) [1, 2] type active immittance functions has been receiving considerable research interest presently. Albeit quite a number of passive tuned L and D immittance synthesis schemes have been proposed in the recent past [3–6], their electronically variable versions are quite few,

as listed in Table 1 [7–17]. The proposed immittance designs utilizes a composite *Multiplication Mode Current Conveyor* (MMCC) [17–19] *Current Feedback Amplifier* (CFA) [20, 21] *active building block* (ABB), as shown in Fig. 1, where simple variation of the nodal d.c.–control voltage of the MMCC device tunes the values of the L and D parameters without requiring any circuit extension complexities. The databook [20] indicates that the CFA has some device imperfections, viz. parasitic capacitances [21–23] at its y and z-nodes ($3.3< C_{y,z}\,(pF) < 6.3$) while its high-frequency roll-off poles are located at relatively high-frequencies around fo$\sim$150 MHz and these are located at close proximity [23, 24]. Effects of these device imperfections have been examined to be quite negligible on the derived values of L and D. Subsequently, functional utility of these immittances had been studied with appropriate design of high-quality(Q) bandpass-LC and band-reject-rD type filters. Previously reported electronically variable immittance function formulations are shown in Table 1.

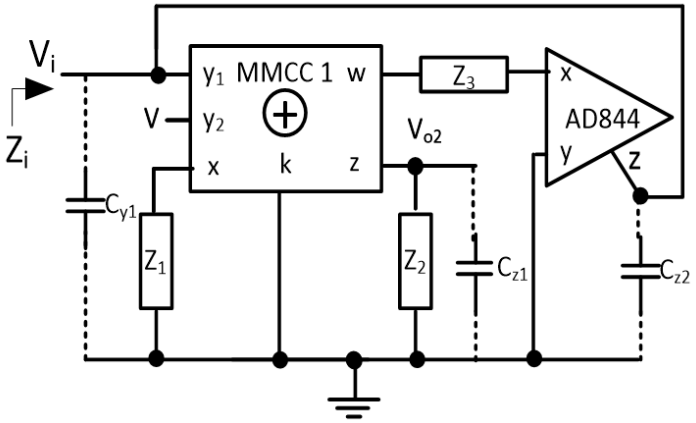


**Fig. 1.** Proposed voltage tunable immittance (VTI).

Hence the new ideas presented in this article may be summarized as: i) Proposition of electronically tunable designs of L and D type immittances based on the new MMCC device ii) The feature of electronic-tunability is derived simply by utilizing the control d.c. voltage (V) available inherently with the MMCC-device. These ideas lead to the main research motivations of this paper, namely: i) derivation of electronically tunable immittance functions by virtue of the control voltage-V available in-situ with the device utilized, and ii) applications of the immittances to the design of selective filters and LC-type linear voltage controlled quadrature oscillator (LVCQO). Voltage tunable immittance functions have earlier been proposed in [17, 18], where the oscillator implementation needs a pair of multiplier elements for linear tunability feature. Here the oscillator linear tuning is simplified by just one multiplier device.

Here a *voltage tunable immittance* (VTI) circuit is designed as described in Section 1, to realize the electronically tunable inductor (L) and *Frequency-Dependent Negative Resistance* (FDNR) (D) type immittances, which are analyzed in Section 2. Parasitic effects in the design circuits are also considered in Section 3. Next, using the VTI, selective filters, viz. Band-pass, Band-reject, are designed and analyzed in Section 4, subsequently voltage controlled (LC-type and DR type) oscillators in Section 5. All experimental results of the filters and oscillators are thoroughly described in Section 6. The attributes of the proposed circuit and its applications are intensively discussed along with supporting experimental results in Section 7.

**Table 1.** Summary of recent immittance functions

| Ref. | ABB Used | Immittance type | Tunability by | Select frequency (MHz) |
|---|---|---|---|---|
| [7] | DO-CCCII | L | $I_b$ | 0.1 |
| [8] | CCTA | D | $I_b$ | 0.14 |
| [9] | DVCCTA | L and D | $g_m$ | 2.5 |
| [10] | VDVTA | L and D | $g_m$ | 1.2 |
| [11] | OTA | L | $g_m$ | 0.2 |
| [12] | VDTRA | D | $\sqrt{I_b}$ | 0.1 |
| [13] | CFA-4 and OTA-2 | L | $\sqrt{I_b}$ | 4.0 |
| [14] | ETDVCCTA | L and D | V | 5.0 |
| [15] | CFOA OTA | C-Multiplier | $I_b$ | 7.0 |
| [16] | Ex-CCCII (using 5-CFAs) | L and D | $\sqrt{I_b}$ | 0.01 (filter response) |
| [17] | CFA-Multiplier | L and D | V | 6.3 |
| [18] | Multiplier-loop | - | V | 13.2 |
| Proposed | MMCC-CFA | D and L | V | 5.1 |

## 2. Immittance Functions

The proposed immittance function realization schemes are shown in Fig. 1; the device node equations are $I_z = \alpha_{1,2}\ I_x, V_w = \delta_{1,2}V_z$ and $V_x = \beta_2 V_y, V_x = \beta_1 kVV_y;\ k \approx$ 1/volt.d.c.[19] where $k$ is the multiplication constant; the subscripts 1 & 2 are for the MMCC and CFA respectively while $C_{z1,2}$ and $C_{y1,2}$ are the device parasitic capacitances ($3.3 < C_{y,z}$ (pF) $< 5.7$) [22]. The port coefficients, assuming device roll-off poles, may be expressed as $\alpha_{1,2} \approx (1 - \varepsilon_{i,1,2})/(s\tau_{i,1,2}+1), \beta_2 \approx (1-\varepsilon_{v,2})/(s\tau_{i,1,2}+1), \delta_1 \approx (1-\varepsilon_{o1})/(s\tau_{i,o1}+1)$. The dc-gain errors are quite low ($\varepsilon << 1$).

The proposed realization circuit is shown in Fig. 1; derivation of the input impedances ($Z_i$) is evaluated as shown in Table-2. These are derived as $Z_i = Z_1Z_3/(kVZ_2)$ assuming ideal ABB elements ($\varepsilon << 1; \alpha \approx \beta \approx \delta = 1$). With appropriate choice of the port RC-components, the proposed immittances are derived as illustrated in Table 2.

The modified input impedance in Fig. 1 is

$$\hat{Z}_i = Z_i/\alpha_2\beta_2\ \ \alpha_1\delta_1 \tag{1}$$

where: $\alpha_{1,2} \approx (1-\varepsilon_{i,1,2})/(s\tau_{i,1,2}+1), \beta_2 \approx (1-\varepsilon_{v2})/(s\tau_{v2}+1)$, and $\delta_1 \approx (1-\varepsilon_{01})/(s\tau_{01}+1)$; since the rolloff poles appear closely at relatively high-frequency-range , we may be assume $\tau_{i,v,z} \approx \tau \equiv 1/\omega_p$ that yields

$$\alpha_2\beta_2\alpha_1\delta_1 \approx 1/\{(s\tau)^4 + (4s\tau)^3 + (6s\tau)^2 + 4s\tau + 1 \tag{2}$$

Neglecting higher order-terms and writing $s\tau = j\omega\tau \equiv j\omega/\omega_p \approx ju$ for the frequency-domain, the modified L-value is obtained, assuming $u << 1$, as

$$\acute{L}/L = \delta L = 1/\sqrt{\{16u^2+1\}}@\llcorner \arctan(4u) \tag{3}$$

**Table 2.** Realization of L and D elements in Fig. 1

| Type | Component selection | Immittance |
|---|---|---|
| a. | $Z_{1,3} = R_{1,3}$<br>$Z_2 = 1/sC$ | $Z_i = sL$ and $L = R_1R_3C/kV \approx CR^2/kV$ |
| *b.* | $Z_{1,3} = 1/sC_{1,3}$<br>$Z_2 = R$ | $Z_i = 1/s^2D \;\; D = C_1C_3R(kV) \approx RC^2(kV)$ |

Hence effects of the device rolloff poles are quite negligible. Similar derivations for the D-element may also be derived, given by

$$\check{D}/D = \delta D = \sqrt{\{}16u^2 + 1\}@\llcorner - \arctan(4u) \tag{4}$$

Hence effects of the device rolloff poles are quite negligible. Similar derivations for the D-element may also be derived, given by

# 3. Effects of Parasitic Capacitances

The effects of device parasitic capacitances are analyzed as

$$\tilde{Y}_{iL} = \{sC_p + (1/s)\tilde{L}\} \tag{5}$$

where $C_p \approx C_{z1} + C_{y2}$. Therefore

$$\tilde{L} = L(1+\mu) \approx L;\; \mu \approx C_{Z2}/C << 1 \tag{6}$$

and

$$\tilde{Y}_{iD} = \{sC_p + s^2\tilde{D}\} \tag{7}$$

where $\tilde{D} = D/(1+jp);\; p = \omega/\omega_{z2};\; \omega_{z2} = 1/RC_{z2}$ .

$$i.e.\; \tilde{D}(\omega) = D/(p^2+1)| \approx D;\; p << 1 \tag{8}$$

The measurement indicates $C_{y1}, C_{z1}, 2 \sim 3.7pF$, *i.e.* with usual R-values in k$\Omega$ range, one gets $f_{z2} \sim$ 57MHz; hence 3-dB rolloff for constant nominal value of D in Eq. (7) appears at quite a higher corner frequency. The useful frequency-range (p<<1) of the FDNR applications has therefore been assumed constant for the proposed design applications here.

# 4. Filter Design Applications

The proposed L and D immittances are next utilized for the design of new LC-type bandpass (BP) and rD-type band-reject (BR) filters as shown in Fig. 2 (a) and (b), respectively. The design equations are summarized in Table 3. Effects of parasitic capacitors ($C_{z,p}$) on stability factor $S_F$ have been investigated as shown in Table 4.

Effects of the port errors ($\varepsilon_{i,v,z} << 1$) on filter circuit parameters *i.e.*, quality factor (Q) and pole frequency ($\omega_o$) are quite negligible, given by

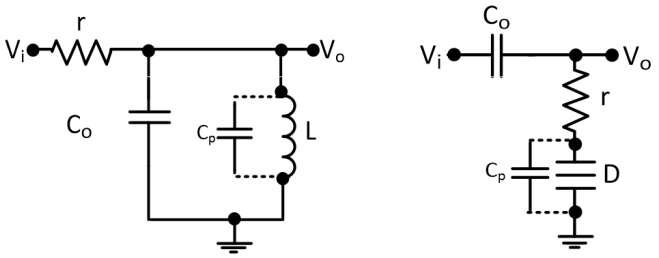


**Fig. 2.** (a) Bandpass filter (b) Band-reject filter

$$Q'_{bp} \approx Q\sqrt{1-\varepsilon_T} \quad \text{and} \quad Q'_{br} \approx \frac{Q}{\sqrt{1-\varepsilon_T}} \tag{9}$$

$$\omega'_{bp} \approx \omega_o\sqrt{1-\varepsilon_T} \quad \text{and} \quad \omega'_{br} \approx \frac{\omega_o}{\sqrt{1-\varepsilon_T}} \tag{10}$$

where $\varepsilon_T = \varepsilon_{i1} + \varepsilon_{i2} + \varepsilon_{v1} + \varepsilon_{o2}$ .

**Table 3.** BP and BR filter design schemes

| Fig. 2 | Filter Transfer Function (F) | Filter parameters | Measured Values |
|---|---|---|---|
| (a) | LC-type BPF : $\mathrm{F_{bp}} = \frac{s\frac{L}{r}}{s^2\mathrm{LC}(1+\sigma)+\mathrm{s}\left(\frac{L}{r}\right)+1}$ where $\sigma = \frac{C_p}{C} << 1$; $\mathrm{C_p}= C_{y1} + C_{z2}$ | $\omega_o = \frac{1}{C}\sqrt{\frac{\mathrm{kV}}{\mathrm{rR}}}$ and $\mathrm{Q} \approx \frac{r}{\omega_o L}$ | $f_o \approx 5MHz$ $Q \approx 4.6$ |
| (b) | rD-type BRF: $F_{\mathrm{br}} = \frac{s^2+\mathrm{s}\frac{C_p}{D}+\frac{1}{\mathrm{rD}}}{s^2+\mathrm{s}\left\{\frac{C_p}{D}+\frac{1}{\mathrm{rC_o}}\right\}+\frac{1}{\mathrm{rD}}\left[1+\frac{C_p}{C_o}\right]}$ $\approx \frac{s^2+\frac{1}{\mathrm{rD}}}{s^2+\mathrm{s}\frac{1}{\mathrm{rC}_o}+\frac{1}{\mathrm{rD}}}$ where $\mathrm{C}_p/\mathrm{D} \approx \frac{C_p}{C^2}/RV^2 << 1$; as $\mathrm{C}_p << \mathrm{C}$ | $\omega_o = \frac{1}{C\sqrt{\mathrm{kVRr}}}$ and $\mathrm{Q} \approx \frac{\mathrm{C_o}}{C}\sqrt{\frac{r}{\mathrm{kVR}}}$ | $f_o \approx$ 4.5MHz $Q \approx 10$ |

Capacitors : $C_{y,z(1,2)\sim}$ 3.6 pF

**Table 4.** Measured Value of Parasitic Capacitors : $C_{y,z(1,2)\sim}$ pF

| Filter Type | Parasitic components | $S_F\|u = \omega/\omega \approx 1$ |
|---|---|---|
| LC-type BPF | Capacitors ($C_{z,p}$) $\sigma = C_p/C_0 << 1$ $\mu = C_{z1}/C << 1$ | $S_F = 2\frac{r}{R}\sqrt{\frac{kVC_o(1+\sigma)}{C(1+\mu)}}$ $\approx 2\frac{r}{R}\sqrt{\frac{kVC_o}{C}} >> 1$ $as \frac{r}{R} >> 1, C \geq C_o$ |
| rD-type BRF | Capacitors (Cp) $\sigma = C_p/C_0 << 1$ $\varphi = C_p/C << 1$ | $S_F = \frac{2\sqrt{\frac{1+\sigma}{rD}}}{\left[\frac{C_p}{D}+\frac{1}{rC_o}\right]} \approx \frac{2}{\varphi}\sqrt{\frac{kVR}{r}} >> 1$ $as \frac{C_o}{C} >> 1, R \geq r$ |

Effects of relative change in o may be expressed [38] as $(\Delta\omega_o/\omega_o) \approx -\{(\partial R/R)+(\partial C/C)\}$for

the change of temperature of 0°C to 100°C, which shows variation of component as $\partial C/C = (-)1.9\%$ and $\partial R/R \leq (+)2.1\%$ (max)[39–40]; hence $\Delta\omega_O/\omega_O \sim 0$.

# 5. Voltage Control Oscillator (VCO) Design

The analysis of the state-of-the-art indicates quite a number of electronically tunable *Quadrature Oscillators* (QO) were proposed recently in [27–38], where only a few exhibits a linear $f_o$-tuning law as listed in Table 5. Some of these are tuned by the device bias current ($I_b$) or transconductance ($g_m$). Next, VTI circuit in Fig. 1 has been modified as shown in Fig. 3, to implement the VCOs by putting input node-$V_i$ at ground. Here the input impedance ($Z_i$) at node $V_o$, derived as $Z_i = Z_1 Z_3/[Z_2(kV)^2]$.

The characteristic equation (CE) of these proposed oscillators in Fig. 3 can be obtained as [37] following the continuous-time model, in terms of

$$Z_L + \mathrm{Z}_i = 0 \tag{11}$$

In (11), choosing $Z_X = 1/sC_o$ and r, in combination with $Z_i = s\widetilde{\mathrm{L}}$ and $1/s^2\widetilde{\mathrm{D}}$ respectively, the CE of LC and rD type Voltage Control Oscillators yields

$$\frac{1}{\mathrm{sC}_O} + \mathrm{s}\widetilde{\mathrm{L}} = 0 \quad \text{and} \quad \mathrm{r} + \frac{1}{s^2\widetilde{D}} = 0 \tag{12}$$

**Table 5.** Summary of electronically tunable QOs

| Ref. | ABB | $f$o (MHz) | Tuning by | Linear Tuning Law | THD% |
|---|---|---|---|---|---|
| [27] | CDTA | 1.73 | $\sqrt{I_b}$ | No | 3 |
| [28] | VDIBA | 8.5 | $\sqrt{I_b}$ | No | 2.2 |
| [29] | DVCCTA | 3.18 | $\sqrt{I_b}$ | No | – |
| [30] | DVCCTA | 1.6 | $\sqrt{I_b}$ | No | 2 |
| [31] | CCTA | 1.7 | $\sqrt{I_b}$ | No | 0.3 |
| [32] | CCCTA | 3.1 | Ib | No | 4.28 |
| [33] | ETDVCCTA | 9 | V | Yes | 3.3 |
| [34] | CFTA | 5.1 | $\sqrt{g_m}$ | No | 2∼4.5 |
| [35] | VTCCTA | 9.9 | V | Yes | 1.5 |
| [36] | CDBA-2 | 0.13 | MOS switch | No | 3.4 |
| [37] | CFA-1 & Multiplier | 12 | V | Yes | – |
| [38] | DX-MOCCII* | 5.62 | MOS bias voltage | No | 3.1 |
| Proposed | MMCC-CFA | $\sim 13.6$ | V | Yes | 3.6 |

Hence these frequency of oscillation (FO), $\omega_{oL}$ and $\omega_{oD}$ for LC-type and rD-type oscillator respectively may be derived using the performance specifications of the oscillation frequencies written as

$$\omega_{\mathrm{oL}} = \frac{1}{\sqrt{\widetilde{L}C_o}} \; approx \frac{\mathrm{kV}}{\sqrt{C_o \mathrm{C} \mathrm{R}_1 R_3}} \tag{13}$$

as $\widetilde{L}$= L(1($\mu$) $\approx$ L =$\frac{\mathrm{CR_1}R_3}{(\mathrm{kV})^2}$ as $Z_{1,3} = R_{1,3}, Z_2 = 1/sC, \mu << 1$.

Similarly, selecting $Z_{1,3} = 1/sC_{1,3}$ and $Z_2 = R_1$,

$$\omega_{\mathrm{oD}} = \frac{1}{\sqrt{\mathrm{r}\widetilde{\mathrm{D}}}} \approx \frac{1}{\mathrm{kV}\sqrt{C_1C_3R_1r}} \tag{14}$$

as $\widetilde{D}$= D(1($p$) $\approx$ D =$\mathrm{R}_1C_1C_3(\mathrm{kV})2$ and $p << 1$

The desired computation of the oscillator-parameters is derived in terms of (13) and (14) as per requirement of the oscillation-frequency range. This computation is exemplified in Section 6.

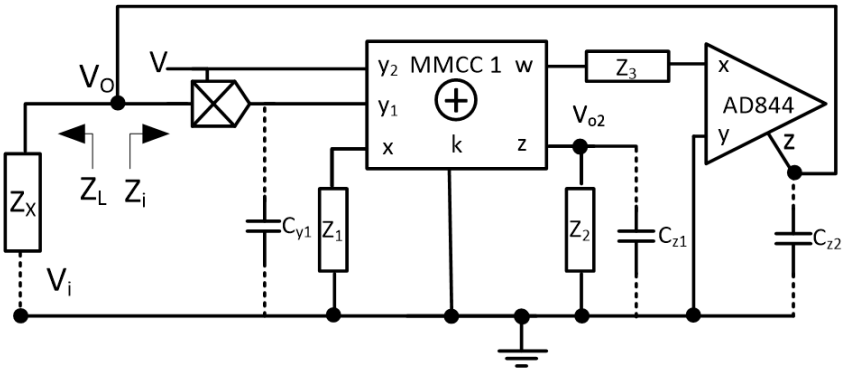


**Fig. 3.** Modified immittance function based Voltage Controlled Oscillator.

Hence the objectives of the control are clearly outlined by Equations (13) and (14) that exhibit the oscillation frequency ($f_o$) is tunable by the multiplier control voltage (V), where the LC-type oscillator yields a linear voltage control law with quadrature property. Several linear and nonlinear controllers with successful applications in various fields and those applications are closely related to the process treated here [44–46].

# 6. Experimental Results

The BP, BR filter and VCO responses are measured from the hardware design with d.c. supply voltage $\pm$5 volt and k=1/V of the MMCC, using thin film technology based passive SMD components to alleviate the temperature sensitivity issue; these are shown in Fig. 4. The bandpass (BP) filter shows its high selectivity response at center frequency $\sim$5 MHz in Fig. 4 (a) with the design value as $r = 2.7\mathrm{K}\Omega$, R=820$\Omega$, $C = C_o = 56$pF at V=2.1 volt d.c. Similarly, the bandreject (BR) filter response has been obtained with the center frequency 4.4 MHz with $r = R = 820\Omega$, $C = 47$pF, $C_o = 470$pF at $V = 0.91$ volt d.c. The THD variation at this pole frequency for a wide range of applied input sinusoidal peak-peak amplitude has been verified for both filters in Fig. 4 (b).

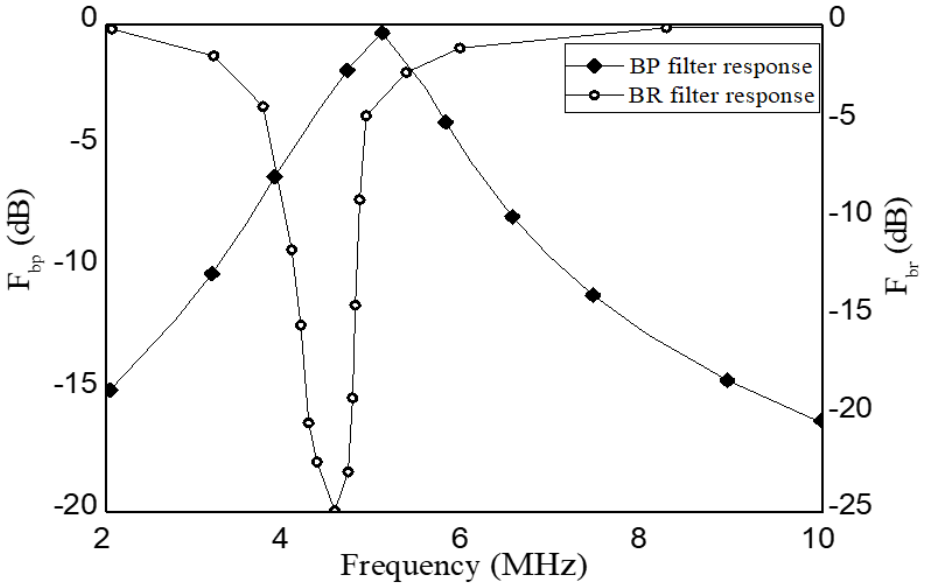


**Fig. 4a.** Filter response of the filter circuit in Fig. 2.

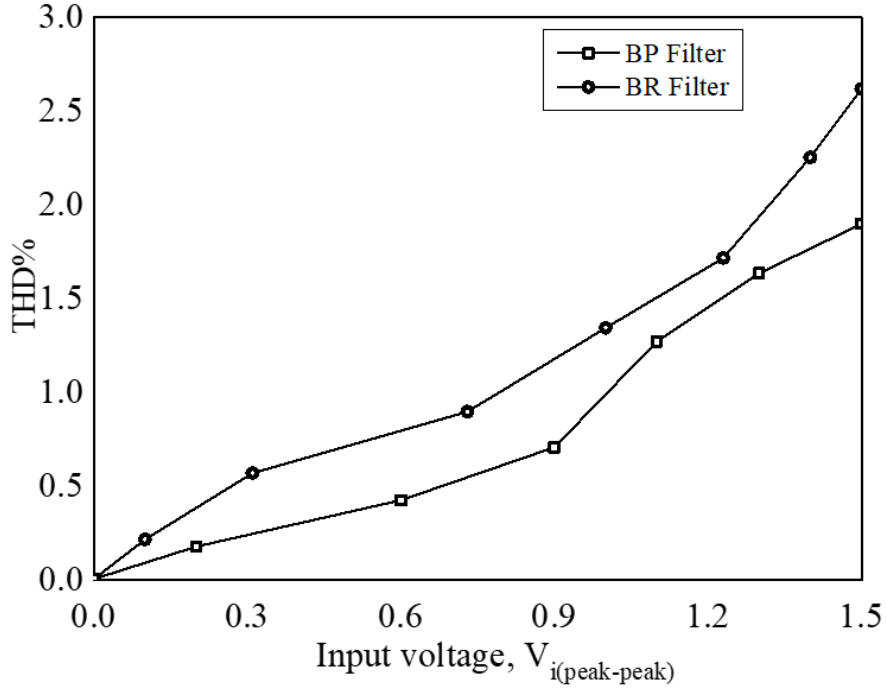


**Fig. 4b.** THD response of the filter circuit in Fig. 2.

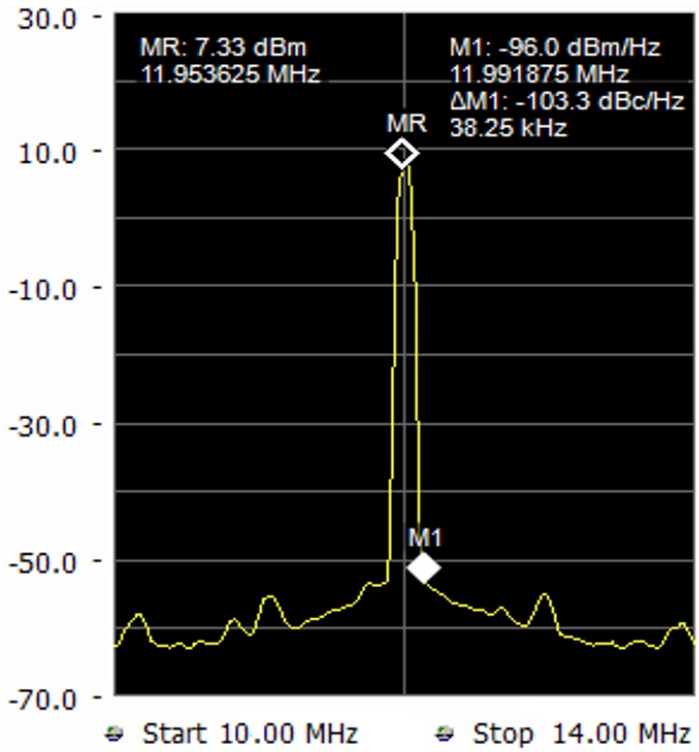


**Fig. 4c.** Spectral response and phase-noise of LC-type linear voltage control quadrature-oscillator.

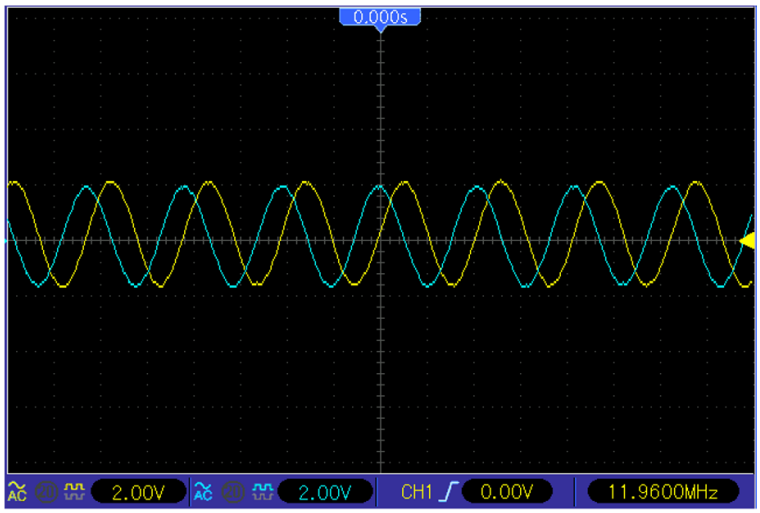


**Fig. 4d.** QO waveform of LVCQO ($V_o$- yellow trace and $V_{o1}$- blue trace) in Fig. 3 .

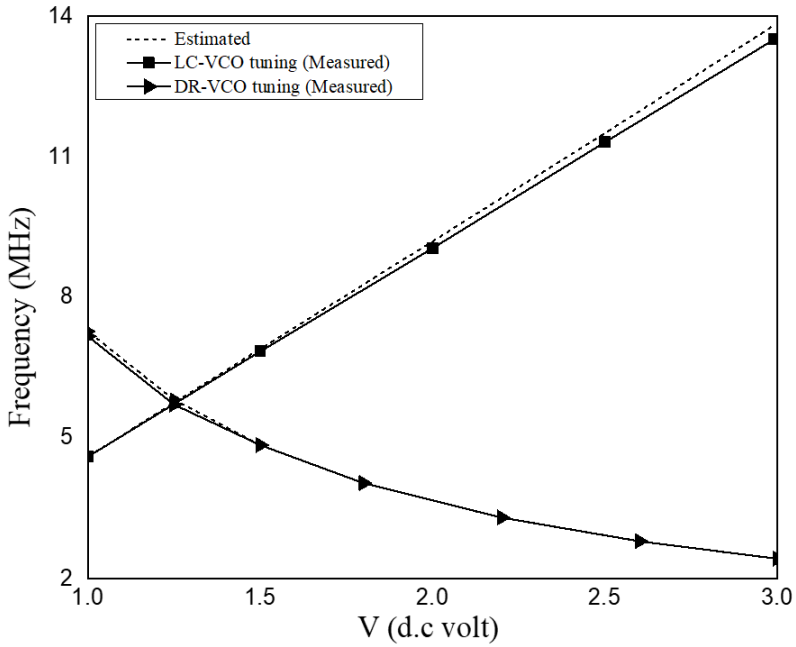


**Fig. 4e.** Tuning law of LC and DR type VCO .

The LC-type VCO response at V=2.6 volt d.c the spectrum at $f_o$ ∼12MHz and its phase noise of (–)103 dBc/Hz at 38 KHz offset at node-Vo has been shown in Fig. 4 (c). Hence an enhanced frequency with improved spectrum quality has been observed in comparison to the [27]. Thereafter its quadrature waveform in between node-Vo and Voq in Fig. 3 at $f_o$ ∼12MHz with 1.5% and 1.7% THD, have been obtained as shown in Fig. 4 (d). The tuning law of the proposed LC and rD-type VCO circuit has been experimentally verified with suitably chosen the values of R1=R3=470Ω, C=Co=74pF and R1=r=470Ω, C1=C3=47pF respectively as depicted in Fig. 4 (e). The linearity error ($\Delta$) in oscillator response had been evaluated following the definition [41]. In Fig. 4 (e), the measured slope of the LC oscillator for the range 4.6MHz ∼ 13.6MHz is (13.56–4.61)MHz/(3–1)Vd.c.= 4.48MHz/V $\equiv \epsilon 2$. Calculated slope $\epsilon 1 = 4.6$MHz/V.d.c. Hence $\Delta = (\epsilon 1 - \epsilon 2)/\epsilon 1 \equiv 2.6\%$.

# 7. Conclusions

A new electronically variable immittance function circuit implementation scheme was presented for the design of electronically tunable L and D type immittances; subsequently measured responses of the frequency selective BP, BR filters along with LC and rD type VCQs are examined. Experimental results on linear QO- response with satisfactory frequency-stability factor and phase-noise figure have been included. Wave harmonic distortion of the quadrature sinusoid signal and the linearity error is quite low for the designed QO. Along with this, a highly selective voltage tunable BP and BR filter with low THD has been obtained. Effects of CFA-device parasitic components are negligible on the frequency-stability ($S_F$) figure.

In the context of the proposed work, of some recent and classical papers [44–46] on several linear and nonlinear controllers with successful applications in various fields, these useful applications are closely related to the process treated in this work. The response of the proposed circuit-designs has been compared with recent research literature that exhibits a minimum component based compact design with better results on the proposed implementation.

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