Mode-Coupling and Phase-Injection Mechanism Enables EMI-Insensitive Crystal Oscillator Circuits

Ulrich L. Rohde\textsuperscript{1}, Ajay K. Poddar\textsuperscript{2}

\textbf{Abstract} – We report a novel crystal oscillator circuit using mode-coupling and phase-injection techniques for improved electromagnetic interference (EMI), start-up, and drive-level, dynamics for the realization of high frequency reference frequency standards. Experimental results and CAD simulated data provides insight into observed characteristics and validated with the 155.6 MHz crystal oscillator (XO).

\textbf{Keywords} – EMI, Crystal Oscillator, CAD, XO

I. INTRODUCTION

Modern communication systems are particularly susceptible to electromagnetic interference (EMI) that can induce the crystal oscillator circuits to oscillate at a different mode and sub-harmonics. Most of the communication systems rely on crystal oscillator as reference oscillators for synthesis of the harmonic signal required for their operation. But mode-jumping phenomena of these reference sources limits the interference susceptibility that affects directly the system performance like, bit error rate (BER) in point-to-point radios, range and load capability of telephone networks, reliability of the navigation systems, and detection ability of radars etc.

Many research works \cite{1-13} have exploited different crystal oscillator circuit architectures, which usually play the important role in improving EMI susceptibility, start-up, drive-level, and negative resistance dynamics of the crystal oscillator circuits. But they are expensive and exhibit inferior phase noise. In this work, we report a low cost solution that partially overcomes the above problem, while maintaining the high stability and phase noise performances.

II. CRYSTAL OSCILLATOR CIRCUIT MODEL

Fig. 1 shows the typical circuit model of a quartz crystal that includes all the modes, where series resonant circuits ($L_i$, $C_i$, $R_i$ with $i=1, 3, 5,...$) represent a mechanical modes (excited by the piezoelectric effect), and $C_0$ is the holder capacitance.

The inductance $L_i$ is the electrical equivalent of crystal mass, capacitor $C_i$ represents the crystal stiffness or elasticity, and resistor $R_i$ represents the heat dissipation due to mechanical friction in the crystal (Fig. 1). In general, crystals are optimized with respect to a particular mode based on the design constraints and performance requirements suitable for typical crystal oscillator topology. The first LCR branch ($L_1$, $C_1$, and $R_1$) models the fundamental mode of oscillation, and the other branches ($L_{n0}$, $C_n$, and $R_n$) are odd overtones, which are calculated as $L_{n0} = L_i$, $C_n = C_i/n^2$, $R_n = n^2R_i$, where $n$ is the $n$th overtone ($n=3$, 5, 7...).

The expression of the impedance $Z(s)$ for a particular resonance condition is given by

\begin{equation}
Z(s) = \frac{1}{sC_0 + \frac{1}{sC + sL_i + R_i}} = \frac{s^2 + s\omega_n^2/Q + \omega_n^4}{sC_0}(s^2 + s\omega_n^2/Q + \omega_n^4) = \frac{(s-s_{1p})(s-s_{2p})(s-s_{1s})(s-s_{2s})}{C_0(s-s_{1p})(s-s_{2p})(s-s_{1s})(s-s_{2s})}
\end{equation}

\begin{equation}
s_{1s,2} = \frac{-\omega_n^2 \pm j\text{\omega}_o}{2Q}\sqrt{1 - \frac{1}{4Q^2}} = -\frac{\omega_n^2 \pm j\text{\omega}_o}{2Q} \text{ for } Q >> 1
\end{equation}

\begin{equation}
s_{1p,2} = \frac{-\omega_n^2 \pm j\text{\omega}_o}{2Q} \sqrt{1 + \frac{C_i}{C_0} \frac{1}{4Q^2}} = \frac{-\omega_n^2 \pm j\text{\omega}_o}{2Q} \left(1 + \frac{C_i}{2C_0}\right)
\end{equation}

\begin{equation}
Q = \frac{L\omega_o}{R_i} = \frac{1}{RC_0\omega_o}
\end{equation}

where $s_{1s,2}$ and $s_{1p,2}$ are zero and poles, $\omega_o$ is the series resonance frequency and $Q$ is the quality factor.

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{fig1.png}
\caption{A typical distributed electrical equivalent circuit of a quartz crystal: (a) Includes all modes, (b) particular mode ($i=1, 3, 5, n$)}
\end{figure}

From (1), series mode resonance of the crystal is represented by a loss resistance in series ($R_i$) with a resonant reactance ($X_i$), whereas, in parallel mode by a resistor ($R_i$) in parallel to a resonant susceptance ($B_i$). One can equivalently represent the series and parallel resonance condition from the imaginary part of the zeroes (3) and poles (5) as

\begin{equation}
\omega_o = \frac{1}{\sqrt{L_iC_i}} \Rightarrow f_i(\text{series}) = \frac{1}{2\pi\sqrt{L_iC_i}}
\end{equation}
\[ \omega_p = \omega_0 \left(1 + \frac{C}{2C_0}\right) \Rightarrow f_p(\text{parallel}) = \frac{1}{2\pi \sqrt{L_1 C_1}} \left(1 + \frac{C}{2C_0}\right) \quad (8) \]

\[ M = \frac{2Q(\omega_p - \omega_0)}{\omega_0} = \frac{2Q(f_p - f_0)}{f_0}, \quad M \text{ is mode separation} \quad (9) \]

From (7), the degree to which crystal oscillator generates a constant frequency \( f \) throughout a specified period of time is defined as the frequency stability of the signal source. But in reality, other modes may exist under the influence of electromagnetic interference (EMI), crystal package parasitic, and crystal’s intrinsic overtone mode features, resulting in induced sub-harmonic (lower than the fundamental frequency) and higher order modes (higher than the fundamental frequency). For brief insights about the EMI-induced sub-harmonic response, Fig. 2 shows the simplified model of the external interference influenced crystal oscillator circuit, where \( L_i, C_i \), and \( R_i \) determine the fundamental mode of oscillation, and the crystal lead inductance \( L_{\text{lead}} \) provides magnetic coupling to external interference. The external interference can be modelled by radio frequency interference (RFI) current source and the equivalent inductor \( L_{\text{int}} \), which is mutually coupled to \( L_{\text{lead}} \) [1]. The EMI induced sub-harmonic response can be described by frequency \( f_2 \) as

\[ f_1 = \frac{1}{2\pi \sqrt{L_1 C_1}} \rightarrow \text{Fundamental} \quad (10) \]

\[ f_2 = \frac{1}{2\pi \sqrt{L_i C_i}} = \frac{1}{2\pi \sqrt{(L_i + L_{\text{int}} + pM)C_i}} \rightarrow \text{Sub-Harmonic} \quad (11) \]

\[ p = \frac{\text{EMI Equivalent Inductance}}{\text{Crystal Lead Inductance}} = \frac{L_{\text{int}}}{L_{\text{lead}}} \quad (12) \]

\[ L = L_i + L_{\text{int}} \pm pM \rightarrow \text{Effective Inductance} \quad (M=\text{coupling factor}) \quad (13) \]

From (11), EMI-induced unwanted sub-harmonic oscillation leads to a failure mode of the crystal oscillator. In addition to this, package parasitic can also generate higher order modes, including odd-order overtone modes.

![Fig. 2. A typical simplified crystal oscillator model with RFI source](image-url)

![Fig. 4. A typical 155.6 MHz mode-coupled phase-injected crystal oscillator with external coupled RF pulse source (inductive coupled)](image-url)
And this effect (shift in frequency) is more prone at high frequency (UHF/VHF) range and it affects dynamic range, selectivity, and sensitivity of a receiver. Therefore, designing a low cost EMI-sensitive crystal oscillator circuit at high frequency (UHF/VHF) is challenging task when the circuit is temporarily exposed to continuous or pulsed RF electromagnetic fields, which creates hysteresis (the new state persist even after the RFI source is removed) and leads to hang-up in digital system [1].

Fig. 5 shows the qualitative analysis of the frequency shift due to EMI but there is no guarantee that the simulated frequency response and phase noise result (Figs. 5 and 6) would hold in general because we did not perform the quantitative measurement with respect to operating frequencies, varying RFI strength (in practical situations, electromagnetic field source may be varying in radiation strength depending upon the distance, nearer or far away), coupling conditions, and scattering from nearby objects (power and ground line traces/devices).

Although, the likelihood of frequency shift in a noisy electromagnetic environment is far higher, this paper describes the design methodology to restrict the frequency shift due to external interference, parasitic and spurious characteristics due to overtone modes without much compromising the oscillator start-up characteristic, oscillator drive sensitivity, oscillator noise factor, and phase noise.

III. DESIGN METHODOLOGY

High frequency crystal resonator is a critical element for reference frequency standards in data communications. Low EMI-sensitive reference signal is a key factor to provide high quality processing of digital systems. And, if we could establish the condition for frequency resonance modes by fundamental, we would be able to suppress EMI-induced failures without being influenced by the unwanted mode-jumping and spurious characteristics due to overtone modes characteristics. The new approach described in this paper, includes the mechanism to select the particular desired resonance mode, optimizing the noise factor, start-up dynamics and drive-level sensitivity functions.

A. Resonant Modes, Drive-Level and Start-Up Dynamics

Crystal resonator exhibits both desired and undesired modes therefore, selection of the desired modes is important for designing the high frequency crystal oscillator circuits. For the high frequency operation, a method is needed to select a particular resonance mode so that other active modes such as: EMI-induced sub-harmonics, parasitic modes, and overtone modes fail to co-exist and sustained oscillations. The simplified technique is to maximize the negative resistance generated from the active device network (Fig. 8) for a given particular mode and must yield positive value of resistance for unwanted modes, including EMI-induced oscillations.

Fig. 7 shows the typical AT cut quartz crystal resonator and its impedance characteristics at fundamental and overtone modes [9]. As the resonance frequency becomes higher, the thickness of crystal resonator becomes very thin and care must be taken to minimize the resonator drive level to avoid breakage of the resonator at high operating frequency. To avoid the breakdown phenomena of the crystal resonator at high frequency, selecting higher series loss resistance \( R_i \) will restrict the level of the current \( I_R \) (Fig. 8) accepted and returned from the crystal but at the cost of poor phase noise. In addition to this, it is difficult to maintain low resonator drive level current \( I_R(\omega) \) without reduction in the magnitude of the negative resistance \( R_n(\omega) \) of conventional Colpitts crystal oscillator circuit (Fig. 8).

![Fig. 5. Simulated frequency response of 155.6 MHz crystal oscillator in the presence of coupled interference source as shown in Fig. 3](image)

![Fig. 6. Simulated phase noise plot of crystal oscillator circuit (Fig. 3)](image)

![Fig. 7. Typical crystal resonator: (a) Quartz resonator, (b) AT Cut crystal thickness at different overtones, (c) Equivalent circuit and (d) Impedance characteristics (fundamental & overtones) [9]](image)
The value of the input impedance $Z_m(\omega)$ (looking into the base of the transistor in Fig. 8) is given by

$$Z_m(\omega) \approx R_m(\omega) + jX_m(\omega) = \left[ \frac{Y_{21}(x)}{\omega C_1 C_2} \right] + \frac{1}{\omega C_1} \left[ \frac{1}{\omega C_2} \right]$$  \hspace{1cm} (14)

From [14], negative resistance and resonator drive level at steady-state is described for circuit shown in Fig. 8 as [13]

$$R_s(\omega) = \frac{Y_{21}(x)}{\omega C_1 (C_1 + C_2) C_2}$$ \hspace{1cm} (15)

$$I_s(\omega) \equiv \frac{2I_f(\omega)}{\rho C_2 R_f(\omega)}$$ \hspace{1cm} (16)

Where

$$Y_{21}(x) = G_m(x) = \frac{qI_m}{kT} \left[ \frac{2I_f(x)}{I_f(x)} \right] = \frac{g_{ds}(x)}{\omega C_1 C_2 M_2(\omega)}$$ \hspace{1cm} (17)

$I_f(x)$ and $I_s(x)$ are the modified Bessel functions of order 0 and 1 respectively. From (15) and (16), resonator current drive level $I_d(\omega)$ can be lowered by increasing the value of feedback capacitor $C_2$ (Fig. 8) for a given current $I_d(\omega)$ but at the cost of reduction in the value of negative resistance $R_s(\omega)$.

In order to maintain the same value of negative resistance $R_s(\omega)$ as required to compensate the loss resistance of the crystal resonator at steady-state oscillation, the value of the feedback capacitor $C_1$ (Fig. 8) has to be reduced. But there is a practical limitation of the minimum value of the $C_1$, which is to be decided by a specified value of the load capacitance of the crystal resonator, including the intrinsic base-emitter capacitor ($C_{b_e}$) of the transistor (Fig. 8) [4-6]. Moreover, drive-level parameter determines the overall nonlinearity, which causes amplitude-frequency effect, and degrades the $1/f$ noise performances [10].

B. Drive Sensitivity Factor Dynamics

The dynamic coefficient of the crystal frequency-drive sensitivity in terms of the fractional change induced as a function of the square current can be described by

$$k_d = \frac{f - f_0}{f_0 \times f_0} \Rightarrow f \rightarrow f_0 \cdot \frac{k_d}{f_0 \times f_0} \times I^2$$ \hspace{1cm} (18)

$$df = 2k_d \times f_0 \times \Delta f \Rightarrow df = \frac{2k_d}{f_0} \times f_0 \times \frac{\Delta f}{I}$$ \hspace{1cm} (19)

where $f_m$ is the modulation frequency, the spectral density of the oscillator output signal phase fluctuation $S_{\phi(f)}$ and amplitude fluctuation $S_{a(f)}$ can be described by

$$S_{\phi(f)} = \left( \frac{df}{f_m} \right)^2 = \delta \phi^2$$ \hspace{1cm} (20)

$$S_{a(f)} = \left( \frac{df}{f} \right)^2 = \delta a^2$$ \hspace{1cm} (21)

AM-to-PM conversion due to resonator drive level sensitivity can be given by ‘γ’ as

$$\gamma = \frac{S_{\phi(f)}}{S_{a(f)}} = \left( \frac{2k_d \times f_0 \times f_0}{f_m} \right)^2 = \left( \frac{2k_d \times f_0 \times f_0}{R_s \times f_m} \right)^2$$ \hspace{1cm} (22)

$$[S_{\phi(f)}]_{\text{open-loop}} = [S_{\phi(f)}]_{\text{closed-loop}} = \left( \frac{1 + \frac{f_0}{2Q_s \times f}}{2} \right)^2$$ \hspace{1cm} (23)

where $V$ is the voltage across the crystal resonator at series resonance, $I$ is the current through the motional arm of the resonator (assuming the static capacitance is anti resonated), $P$ is the power dissipated in the resonator ($P = \overline{Q_R}$), $R_s$ is the resonator series resistance, and $Q_s$ is the resonator loaded $Q$ factor. For $f_s / (2Q_s f) > 1$, the value of $k_d$ that results in the equal levels of oscillator closed PM noise level due to AM noise and open loop PM noise can be described by [12]

$$k_d^2 = \left[ \frac{S_{\phi(f)}}{S_{a(f)}} \right]_{\text{open-loop}} \times \left( \frac{(10^4 \times R_s)^2}{(4 \times Q_s \times P)^2} \right)$$ \hspace{1cm} (24)

From (24), $k_d$ that would result in oscillator phase noise due to AM-FM conversion equal to the that due to the conversion of open loop PM noise to closed loop FM noise. For low phase noise applications, values of $k_d$ has to be dynamically optimized by incorporating mode-coupling and phase-injection techniques.

In this paper we describe drive-level optimization techniques to enable high frequency operation (without witnessing breakdown phenomena of the crystal resonator in UHF/VHF frequency range). This can be accomplished by using mode-coupling technique (tuning the $L = C_m$ at higher order modes) and optimizing the ratio $C_{v2}/C_{v1}$ for reducing the start-up time (time interval required for an oscillator to sustain stable output at desired frequency) and drive-level (Fig. 9). Fig. 10 shows the simulated plots of the start-up characteristics (for a given $C_{v2}/C_{v1} = 1.5$ and $C_{v2}$ shorted), we found that there is a trend in reduction of the start-up time we provide the ratio of the coupling capacitors $C_{v1}$ and $C_{v2}$ for a given drive-level $I_d(\omega)$.

![Fig. 8. A simplified crystal oscillator including noise contributions](image)

![Fig. 9. A typical crystal oscillator including noise contributions](image)
Therefore, oscillator noise factor dynamics is the important parameter for the realization of low phase noise crystal oscillator for a given drive-level and start-up time.

C. Oscillator Noise Factor Dynamics

The expression of the oscillator noise factor \( F \) and the phase noise \( \xi(\omega) \) in dBc/Hz for the crystal oscillator circuit shown in Fig. 9 can be described by [11, pp. 132, 180-181]

\[
F = \frac{Y_{21}C_1C_2}{(C_1 + C_2)C_t} \left( \frac{1}{2r_e} + \frac{1}{r_f} \right) \left( \frac{1}{Y_{21}C_1C_t} + \frac{1}{2} \right) \quad (25)
\]

\[
\xi(\omega) = 10 \log \left( \frac{4\pi k T R}{2\pi Q_0^2 \omega^2} \frac{Y_{21}C_1C_2}{(C_1 + C_2)C_t} \left( \frac{1}{2r_e} + \frac{1}{r_f} \right) \left( \frac{1}{Y_{21}C_1C_t} + \frac{1}{2} \right) \left( \frac{1}{\omega_c^2} + \frac{1}{Q_c^2} \right) \right) \quad (26)
\]

\[
\beta^* = \frac{Y_{21}^*}{Y_{21}} \left( \frac{C_1}{C_2} \right)^y \quad \xi_{opt} = \left( \frac{C_1}{C_2} \right)^y \quad (27)
\]

where \( Y_{21}, Y_{21}^* \) is the large signal [Y] parameter of the active device, \( K \) is the flicker noise coefficient, \( AF \) is the flicker noise exponent, \( R_L \) is the equivalent loss resistance of the tuned resonator circuit, \( I_e \) is the RF collector current, \( V_{ce} \) is the RF collector voltage, \( C_1, C_2 \) is the feedback capacitor (Fig. 9). \( Q_u \) and \( Q_l \) are unloaded and loaded Q factors, \( p \) and \( q \) are the drive level dependent constant across base-emitter of the device.

From curve-fitting attempts, the following values for \( p \) and \( q \) in (27) were determined \(( p = 1.3 \text{ to } 1.6; \ q = 1 \text{ to } 1.1 \)). From (26), the expression of the phase noise is [11, pp. 181]

\[
\xi(\omega) = 10 \log \left( k_o + \left( \frac{k_i}{k_0} \right)^y \frac{m^2(1+y)^2}{y^2(y^2+k)} \right) \quad (28)
\]

Where \( k_o = \frac{k T R}{\omega^2 \omega_c^2 V_{ce}^2}, \ k_i = \frac{q^2 \omega_c^2 V_{ce}^2}{4\pi Q_0^2 \omega^2 \omega_c^2 V_{ce}^2}, \ k_k = \frac{Y_{21}^*}{Y_{21}} \left( \frac{C_1}{C_2} \right)^y \), and \( m = \frac{Q_u}{Q_{loaded}} \) are constant for a given drive level with \( y = \frac{C_1}{C_2} \). Differentiating (28) with respect to \( y \) will give condition for \( y_{opt} \) for optimum noise factor \( F_{opt} \) and minimum phase noise for a given bias and device parameters.

From (25) and (27), feedback parameters \( C_1 \) and \( C_2 \) are dynamically tuned for optimum noise factor \( F_{opt} \) for a given values of \( Y_{21} \) and \( \beta_{opt} \), which is dependent upon the drive level across base-emitter of the device. Fig. 13 shows the typical CAD simulated phase noise plot for a 155.6 MHz crystal oscillator circuit (Fig. 4) at offset 1 kHz from the carrier with the carrier with respect to \( y_{opt}, F_{opt} \) and \( y_{opt} \).
From (25) and (28), for different values of noise factor $F$ ($F_3 > F_2 > F_1$), the phase noise can be lowest for $m_{\text{opt}} \approx 0.5$ and $y_{\text{opt}} \approx 2.2$ (regime: $1.1 < y_{\text{opt}} < 3.2$). It can be seen from Fig. 13, for $F_{\text{OPT}} = F_1(\text{dB}) = 6$ dB, phase noise curve shows two minima for the values of $y_1_{\text{opt}} \approx 2.2$ and $y_\text{2 opt} \approx 3.2$, and one maxima corresponding to $y_\text{2 opt} \approx 1$.

### D. Crystal Oscillator Phase Noise Dynamics

There are mainly three noise sources that mostly contribute to the oscillator phase noise: shot noise, thermal noise, and flicker noise. Shot and thermal noise (broadband noise) generates amplitude and phase modulation of the carrier signal, resulting in equally divided amplitude modulation (AM) and phase modulation (PM) noise independent of frequency, and basically sets the noise floor of typical phase noise spectrum. The frequency flicker of an oscillator that appears as a $1/f$ noise spectrum. The frequency flicker of an oscillator that frequency, and basically sets the noise floor of typical phase noise spectrum.

From (26), the modified Leeson equation by introducing $Q_0$, $f_0$, $f_m$, $f_m$, $Q_0$, $Q_0$, $f_0$, $k$, $T$, $P_o$, $R$, $m$, $Q$, $m$, and $m$ are the ratio of the sideband power in a 1Hz bandwidth at $f_m$ to total power in dB, offset frequency, flicker corner frequency, loaded $Q$, unloaded $Q$, noise factor, Boltzmann’s constant, temperature in degree Kelvin, output power, equivalent noise resistance of tuning diode, voltage gain and ratio of the loaded and unloaded $Q$. The expression of dynamic loaded $Q$ is

$$Q_L = \frac{\partial \phi}{\partial f}$$

From (19) and (33), phase noise performance depends on the noise factor $F$ of the oscillator circuit for a given $Q$ factor, therefore, the noise factor $F$ with $m_{\text{opt}} \approx 0.5$ shall lead to improvement in phase noise performance.

### IV. DESIGN EXAMPLE AND VALIDATION

The new approach includes the mechanism to improve the phase noise performances, including minimization of frequency shift due to EMI and overtone modes. The novel crystal oscillator (XO) circuits are designed for reliable performance over temperature from -40 to +85°C and are well suited as a low cost reference frequency standard for applications in industrial, military, and commercial systems.

Fig. 14 shows the typical schematic of the mode-coupled phase-injection locked 155.6 MHz crystal oscillator (XO) circuit, which sets up optimum and noise impedance transfer function by dynamically controlling mode-coupling and phase-injection for a given drive level and start-up time. As shown in Fig. 14, the higher order mode is coupled through output path and feedback to the point where frequency-drive sensitivity of the crystal resonator shows maximal group delay and faster slew rate, resulting, improved phase noise performance with reduced drive-level $I_D(\omega)$.

![Fig. 13](image1.png)

**Fig. 13.** CAD simulated phase noise plot of 155.6 MHz XO (Fig. 4)

![Fig. 14](image2.png)

**Fig. 14.** Schematic of 155.6 MHz mode-coupled phase-injection locked crystal oscillator (XO) circuit (Crystal resonator $Q \equiv 60,000$)
The circuit layout is fabricated on low-loss 62-mil-thick dielectric material with dielectric constant of 3.38. The active device is a discrete low-noise transistor SiGe HBT device.

Fig. 15 shows the CAD simulated phase noise plots of 155.6 MHz crystal oscillator (XO) circuit (Fig. 14), which depicts the 10-15 dB improvement in the phase noise performances. Fig. 16 shows the measured phase noise plot, which agree with the simulated result within 5 dB for three cases: without mode-coupling, with mode-coupling, and with mode-coupled and phase-injection mechanism. The circuit operates at 5V, 30 mA and also suitable for tunable crystal oscillator for VCXO (voltage controlled crystal oscillator) applications.

As shown in Fig. 15, at lower offset (<10 Hz), improvement in phase noise performance is limited due to the influence 1/f noise, which can be optimized by selecting active device (transistor) and crystal resonator that has low value of the flicker (1/f noise) contributions at UHF/VHF ranges.

By dynamic optimization of the resonator frequency drive sensitivity factor (as discussed in section II), influence of 1/f noise can be minimized for high frequency crystal resonator oscillators. The reported work offers cost-effective solution and can be applied for a crystal resonator (both high-Q and low-Q) based VCXOs for substantial reduction in phase noise. The drawbacks of this approach are large real estate area (Printed circuit board size) and complex circuit require for the fabrication of EMI insensitive layout at UHF/VHF ranges.

The biggest challenge is the characterization of EMI simulated interference sources for the validation of EMI-induced failure due to frequency shift caused by induction of sub-harmonics into crystal resonator. In addition to this, phase-injection locking mechanism is very critical, slight variation in phase may cause oscillation to cease, therefore, phase-locked-loop (PLL) option can be more practical for commercial applications.

Although, mode-coupling phase-injection restricts the frequency drift to the greater extents still crystal oscillator can induce sub-harmonic response under the influence of external radio frequency interference (RFI) around the crystal oscillator circuit.

By proper shielding and making arrangement of EMI insensitive traces for power and ground line can improve the stability against EMI-induced failure. Experimental results and CAD simulated data (Section II and III) provides insight into observed characteristics and validated with the 155.6 MHz crystal oscillator circuit.

V. CONCLUSION

This work offers a novel crystal oscillator circuit using mode-coupling and phase-injection techniques for improved electromagnetic interference (EMI), start-up, and drive-level, dynamics for the realization of high frequency reference frequency standards. Experimental results and CAD simulated data provides insight into observed characteristics and validated with the 155.6 MHz crystal oscillator circuit.

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