Feedback Isolation by Piezoelectric Transformers: A Feasibility Study

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Abstract - Signal isolation is needed in power electronics systems that include separate primary and secondary 'grounds'. The feasibility of using a piezoelectric transformer as a galvanic barrier was investigated in this study theoretically and experimentally. The research included the issues of drive, demodulation, bandwidth and common mode rejection. It was found that a small size transformer (2x3mm) could be used effectively to transfer feedback signals from the isolated output of a DC-DC converter to the primary side with a bandwidth of 7kHz when a carrier of 350kHz is used. Wider signal bands can be achieved by applying higher carrier frequencies.

I. INTRODUCTION

Signal isolation is needed in power electronics systems that include separate primary and secondary 'grounds' [1-4]. The general representation of a typical power converter with primary to secondary isolation includes a power stage, a modulator, and an isolation barrier for the feedback signal (Fig. 1a). To ease the accuracy and stability requirements of the signal isolator, the configuration of Fig. 1b is normally preferred. In this case the isolator has to carry the error signal and not the output signal. Comparing the transfer functions of the two configurations we obtain:

$$\frac{V_{out}}{V_{ref}}(Fig1a)=\frac{H_m(\omega)\cdot H_p(\omega)}{1+K\cdot H_m(\omega)\cdot H_p(\omega)\cdot H_{iso}(\omega)} \qquad (1)$$

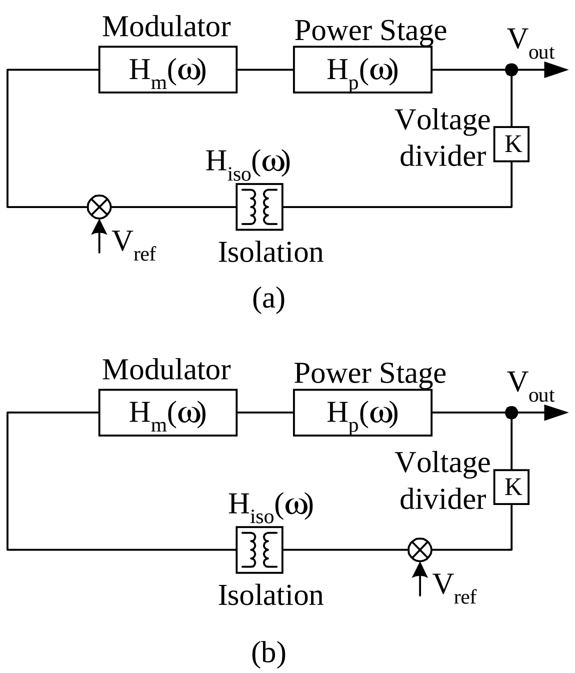

Fig. 1. Alternative feedback loop configurations in an isolated converter. (a) Isolating of output signal. (b) Isolating of error signal.

$$\frac{V_{out}}{V_{ref}}(Fig1b)=\frac{H_m(\omega)\cdot H_p(\omega)\cdot H_{iso}(\omega)}{1+K\cdot H_m(\omega)\cdot H_p(\omega)\cdot H_{iso}(\omega)} \qquad (2)$$

where:

$H_p(\omega)$ - transfer function of power stage

$H_m(\omega)$ - transfer function of modulator

$H_{iso}(\omega)$ - transfer function of the isolator
K – voltage divider
V_{ref} – reference voltage
V_{out} – output voltage

For a large Loop Gain (LG), equations (1, 2) reduce to:

$$\frac{V_{out}}{V_{ref}}(Fig.1a_{LG\to\infty}) \approx \frac{1}{K \cdot H_{iso}(\omega)} \qquad (3)$$

$$\frac{V_{out}}{V_{ref}}(Fig.1b_{LG\to\infty}) \approx \frac{1}{K} \qquad (4)$$

It is thus evident that by choosing the configuration of Fig. 1b (including the voltage reference and error amplifier at the secondary) the system becomes practically independent of the transfer function of the isolator $H_{iso}(\omega)$ – provided that the LG is kept high.

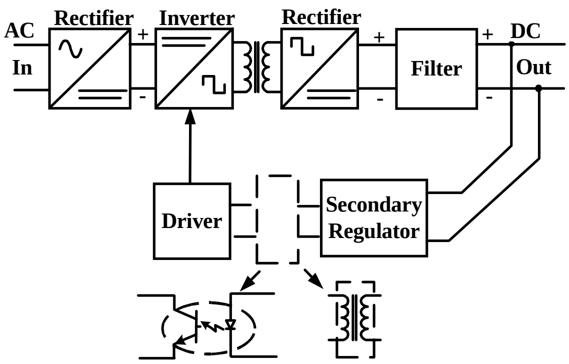

Fig. 2. Conventional feedback isolation in power converters.

Conventional solutions to the signal isolation problem include galvanic isolation by opto-coupler or transformers [1-6] (Fig. 2). Other solutions (such as capacitor coupling [2,3]) are possible, but not all are compatible with interfering signals of high dV/dt normally associated with the power conversion environment. A typical practical design for a flyback converter is shown in Fig. 3. In this case the reference voltage and error amplifier are realized by TL431 (Texas Instruments Inc.), whereas the actual isolation (of the error signal) is carried out by the opto-coupler 4N15.

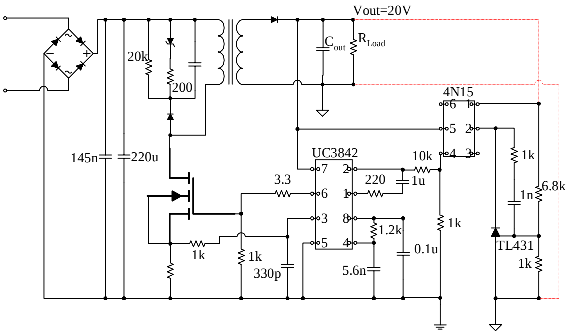

Fig. 3. Flyback converter with opto-coupler isolation.

In this study we explored the feasibility of using a ceramic Piezoelectric Transformer (PT) [7-9] as the barrier element. That is, the possibility of replacing the isolation element (e.g. 4N15 in Fig. 3) by the PT. Since the PT can not transfer DC signal, modulation techniques need to be applied. Namely, the error signal is first used to modulate a carrier and then recovered by demodulating the output signal past the isolation barrier. This is similar to the method used when an electromagnetic transformer is used to realize the isolation barrier [2,3]. The investigation probed into the theoretical aspects of such coupling and included experimental examination of a specific design that uses FM modulation.

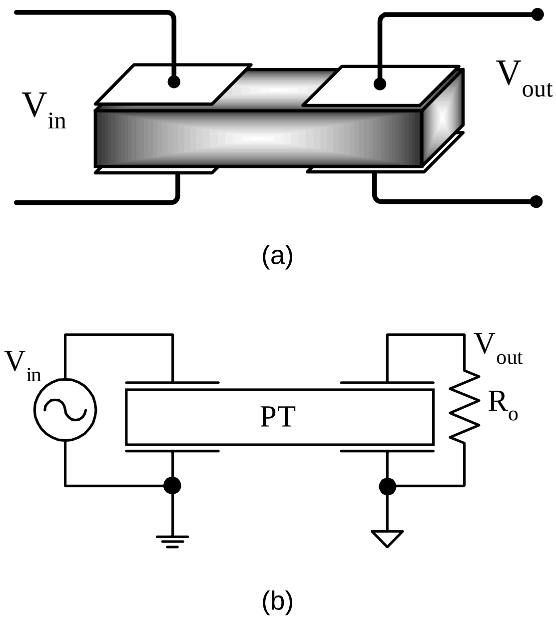

Fig. 4 PT (a) and its connection in the circuit (b).

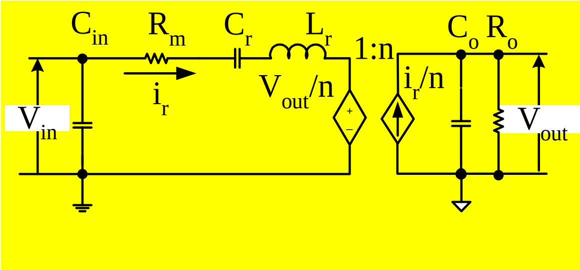

Fig. 5. Equivalent circuit of a piezoelectric transformer (PT) loaded by R_o.

A typical PT is shown pictorially in Fig. 4. The equivalent circuit of a PT (Fig. 5) includes a resonant network (L_r, C_r, R_m) that emulates the effect of the mechanical vibration and dependent sources that express the gain [7-9]. The model also comprises the physical dielectric capacitors (C_{in}, C_o) that are formed by the input and output electrodes. Since the network is highly selective it will pass, with reasonable gain, only frequencies that are in the vicinity of the resonant frequency. The equivalent circuit of Fig. 5 represents one vibration mode but in reality, many such circuits are in fact connected in parallel since the PT can vibrate in various modes. That is, a typical transfer function of a PT will look like the plot of Fig. 6. For a typical application one resonant peak is selected and the design is done for that specific carrier frequency [7].

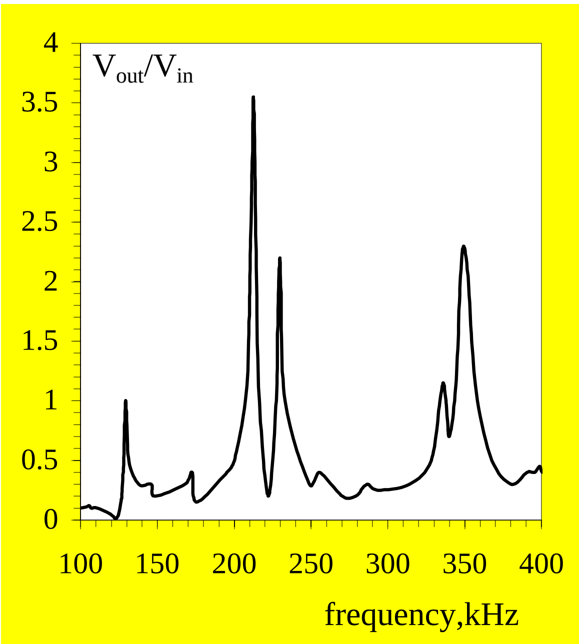

Fig. 6. Gain versus frequency of experimental PT.

II. MODULATION SCHEMES

In this study we explored two modulation alternatives: AM modulation and FM modulation. In the AM case, a constant frequency carrier is used and by varying the amplitude of the carrier the error signal is transmitted. In the FM case, the error signal is coded into a frequency shift that translates at the output of the PT as an amplitude shift. This is illustrated in Fig. 7 that assumes an operating point above the resonant frequency.

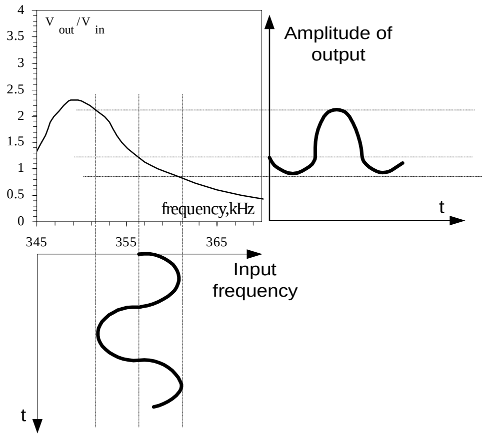

Fig. 7. PT's response to an FM modulated signal.

AM Modulation

Considering the high quality factor (Q) of practical PTs the carrier frequency has to be around the resonant frequency of the device. To explore the effect of modulation it would be desirable to simplify the electrical equivalent circuit of the PT. This can be done by first reflecting the load R_o and C_o (Fig. 8a) to the primary and then converting the parallel $R_o'C_o'$ to a series network (Fig. 8b).

The transformation equations are:

$$
\begin{aligned}
C_o' &= C_o \cdot n^2 \\
R_o' &= \frac{R_o}{n^2} \\
R_o'' &= \frac{R_o}{n^2(1+\omega^2 C_o^2 R_o^2)} \\
C_o'' &= \frac{(1+\omega^2 C_o^2 R_o^2)n^2}{\omega^2 C_o R_o^2} \\
\left|Z_0''\right| &= \sqrt{R_o''^2 + \left(\frac{1}{\omega C_o''}\right)^2} = \frac{R_0}{n^2\sqrt{1+\omega^2 C_o^2 R_o^2}}
\end{aligned}
\tag{5}
$$

where:
n is transformation ratio

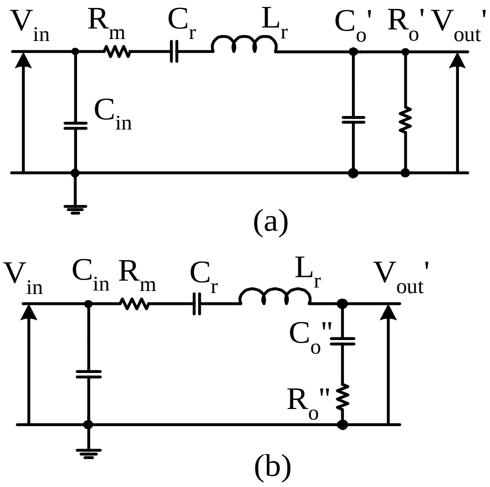

Fig. 8. The stages in simplifying the equivalent circuit of a PT. (a) - reflecting the R_oC_o network to the primary. (b)-translating the parallel R_o'C_o' network into a series network R_o"C_o".

The series elements can now be lumped into an equivalent capacitor C_{eq} and equivalent resistor R_{eq}:

$$R_{eq} = R_m + R_o''$$
$$C_{eq} = \frac{C_r C_o''}{C_r + C_o''} \tag{6}$$

Thus the PT equivalent circuit is reduced into a series RLC network (Fig. 9).

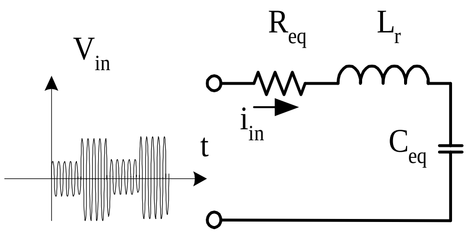

Fig. 9. The simplified equivalent circuit of a PT circuit fed by an AM modulated signal.

Fig. 9 implies that the response of the PT to an AM modulated signal is in fact a response of a series resonant network. An AM signal will be composed of a carrier (f_0) plus two side-bands located at $f_0 + f_s$ and $f_0 - f_s$ where f_s is the modulating frequency. If the carrier is set at the resonant frequency of the PT then the width of the resonant curve (Δf) will be:

$$\Delta f = \frac{f_0}{Q_e} \tag{7}$$

where:
Q_e is the quality factor of the series resonant circuit $Q_e = \frac{2\pi f_0 L_r}{R_{eq}}$ and

f_0 is the resonant frequency of the network $f_0 = \frac{\omega_0}{2\pi} = \frac{1}{2\pi\sqrt{L_r C_{eq}}}$

The bandwidth of a resonant circuit is Δf_{BW} corresponding to the cut off frequency of the modulated signal (f_{so})

$$\Delta f_{BW} = 2 f_{so} \tag{8}$$

from which

$$f_{so} = \frac{f_0}{2Q_e} \tag{9}$$

Hence, by adjusting Q_e (loading properly) one can control the useful signal bandwidth of the PT.

The simplified analysis given above is supported by a rigorous analysis carried out in this study. It was found that the response of a PT to a step function in the carrier can be expressed as:

$$V_{out}(t) =$$
$$n\eta\sqrt{1+C_o^2R_o^2\omega_o^2}\,(V_m(1-e^{\left(\frac{-\omega_o t}{2Q_e}\right)})+V_n)\sin(\omega_o t) = \tag{10}$$
$$n\eta\sqrt{1+C_o^2R_o^2\omega_o^2}\,(V_m(1-e^{\left(\frac{-t}{\tau}\right)})+V_n)\sin(\omega_o t)$$

where:

$\tau = \frac{2Q_e}{\omega_o}$ - is the aparent time constant of the system

$\eta = \frac{R_m}{R_m + R_o''}$ - is the efficiency

V_m is the amplitude of the modulation step

V_n is the amplitude of the carrier

And the bandwidth is:

$$BW = \frac{1}{2 \cdot \pi \cdot \tau} = \frac{\omega_o}{2 \cdot Q_e \cdot 2 \cdot \pi} = \frac{f_o}{2 \cdot Q_e} \quad (11)$$

FM Modulation

Following the same intuitive reasoning as above one can find the useful signal bandwidth in the case of FM modulation. Assuming that a constant frequency shift per unit amplitude of the modulating signal is Δf_m, the useful bandpass will be Δf_m. That is, most of transmitted energy is locked within $f_0 + \Delta f_m$ and $f_0 - \Delta f_m$. In this respect the behavior is similar to the AM case except that the energy is distributed within the range (rather than having only two side bands as in the case of AM). Consequently, the break point of the signal transfer function is reached in this case when:

$$\Delta f_m = \frac{f_0}{2Q_e} \quad (12)$$

Furthermore, the signal bandwidth f_{so} will also be limited to Δf_m since modulating frequencies higher than Δf_m will be highly attenuated. It is thus clear that in this case as in the case of AM modulation the useful bandwidth can be controlled by adjusting Q_e.

III. COMMON MODE REJECTION

A major concern in isolated feedback signal of power systems is the injection of common mode signal via the isolator [1-3]. Such an injection will increase the EMI signal at the output section and will require further filtering to satisfy common standards.

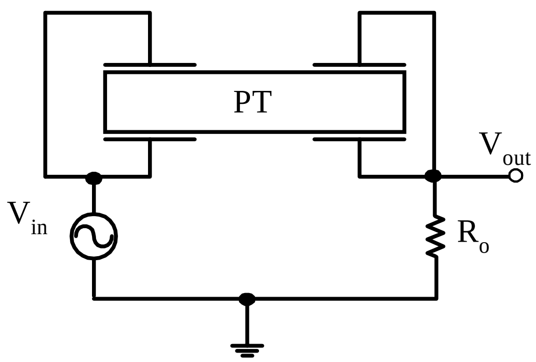

Fig. 10. Setup for measuring common mode transfer ratio.

We have therefore tested the experimental PT for both differential mode gain (Fig. 4b) and common mode gain (Fig. 10). The results (Fig. 11) suggest that the peak at 350 kHz is a good choice since it is a conveniently high frequency and has a large common mode rejection ratio (the ratio of differential gain to common mode gain).

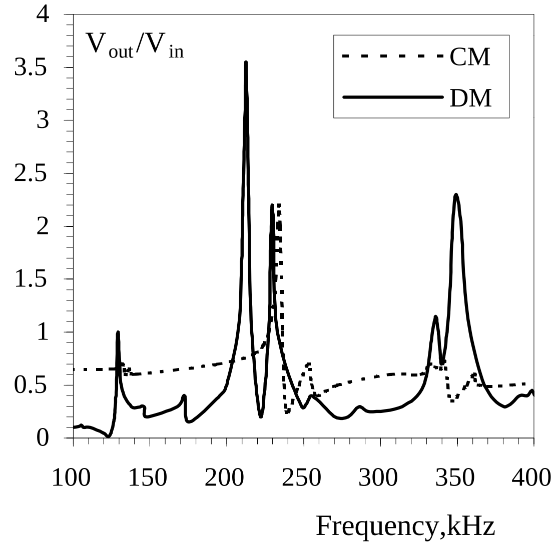

Fig. 11. Differential Mode (DM) and Common Mode (CM) transfer ratios of experimental piezoelectric transformer.

IV. EXPERIMENTAL

The experiment circuit included a small (2x3mm) PT transformer made of PXE43 material (Philips).

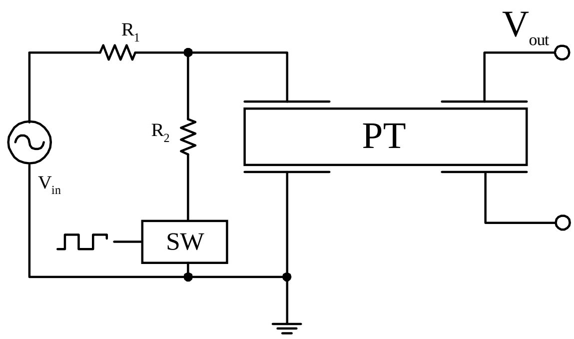

Fig. 12. Setup for measuring BW.

We have used the circuit of Fig. 12 to verify the analysis of signal bandwidth in AM modulation. The element was driven by a 350kHz carrier and the amplitude was changed stepwise by a divider that was controlled by a switch. The results confirm the estimates presented above. The block diagram of Fig. 13 describes the system that was used the PT isolator concept applying FM modulation. The circuit was built around a commercial resonant controller (MC34066). The output was rectified and filtered to recover the error signal. A complete circuit diagram is given in Fig. 14. The design was made compatible to the flyback converter of Fig. 3. That is, the PT isolation system can replace the TL431 and 4N15. The input was designed to be connected to the output voltage while the output of the PT isolator is compatible with the UC3842. During testing, the PT isolation block was fed by a DC signal on which a low frequency AC signal was superimposed.

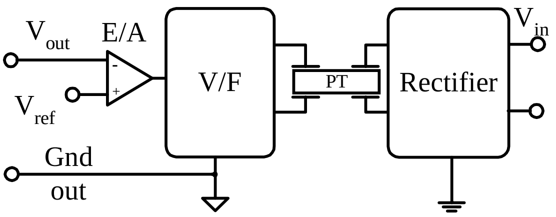

Fig. 13. Block diagram of the experimental PT isolator.

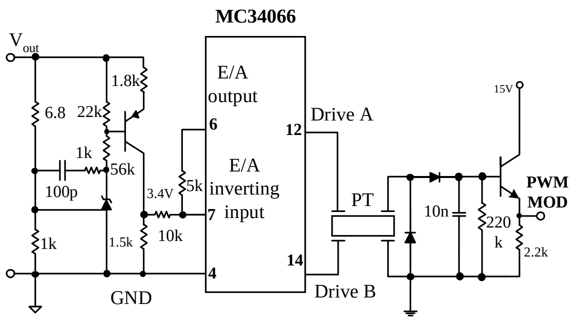

Fig. 14. View of feedback path the whole (practical).

The measured static transfer-function of the isolator reflects the high gain of the error amplifier in MC34066 (Fig. 15). In actual use, the output signal will be forced to approach the reference signal by the global feedback action.

A network analyzer was used to test the signal bandwidth of the FM approach. The measured bandwidth of the experimental circuit was found to be 7KHz (Fig. 16).

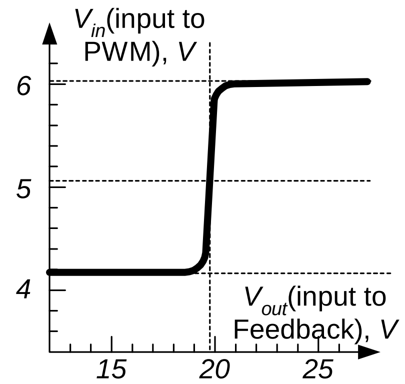

Fig. 15. Static transfer function of the experimental PT isolator.

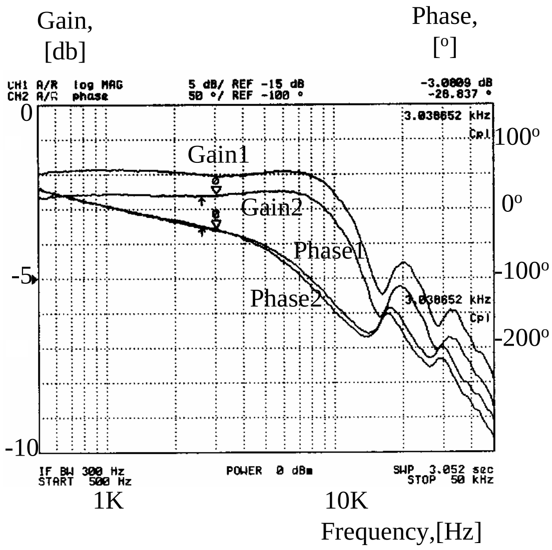

Fig. 16. Transfer function of the experimental PT isolator at two carrier frequencies (gain 1: 358kHz; gain 2: 357.6kHz).

V. DISCUSSIONS AND CONCLUSIONS

The main advantages of the PT in the proposed application are the small size and the very high isolation breakdown voltage that can be achieved [7-9]. This is due to the good isolation of the ceramic material. The main disadvantage is the common mode stray capacitance between primary and secondary sides of the PT. Common mode rejection can be improved by ensuring that the harmonics of the switching frequency of the converter do not coincide with he common mode peaks of the PT.

To fully exploit the engineering benefit of the proposed isolation approach, there would be a need to develop a dedicated IC that will be compatible with the PT. A desirable feature of such an IC would be a self oscillation and locking to the preferred resonance frequency of the device. In the case of AM modulation, locking to the resonant frequency is preferred. In the case of FM, a phase lock loop can be used to stabilize the operating frequency.

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