ABSTRACT In this study, we perform an analytical investigation of the electro-optic (EO), Mach-Zehnder modulator (MZM) transmission boundary area with a nonlinear model. We propose a nonlinear model that represents the allowable RF input amplitude within the MZM transmission boundary area. We investigate the maximum RF input amplitude and confine its harmonics using nonlinear model theory, RoF simulation, and MZM characteristic experiments. The maximum RF input amplitude in the MZM transmission boundary area is a convenient technique that can confine harmonics without an optical amplifier, optical filter, or other optical devices in RoF links. We select several RF input amplitudes within the MZM boundary area and account for harmonic distortion in the investigation. The RF input amplitude is 0.8 times of the EO modulator switching voltages. The boundary areas show that near the switching voltage value, the harmonics-to-main RF ratio (HMR) is less than 10%. Furthermore, the maximum RF inputs are still appropriate for QPSK modulation. We evaluate the nonlinear model of these modulation signals in terms of the error vector magnitude and constellation diagram. After considering the optical fiber impairment, thermal noise, and harmonics distortion in the nonlinear model with the modulation signal of QPSK, the ACLR and EVMs satisfy the 3GPP specification. The nonlinear models are appropriate for a small or medium sized of radio-over-fiber links.

INDEX TERMS Radio-over-fiber, electro-optic modulator, MZM transmission boundary area, RF input, harmonics, error vector magnitude, analytical and experimental investigation.

I. INTRODUCTION

Integrated antenna-EO modulator devices are desirable for future technological applications [1], as they reduce the need for EO conversion power, decrease signal latency, obtain the entire EO conversion process in the RoF system and reduce device sizes [2], [3], [4], [5], [6], [7], [8], [9], [10], [11], [12], [13], [14], [15], [16]. These devices feature antennas that convert radio frequency (RF) signals directly by using patches, and an external modulator electrode performs EO conversion. The optical materials in these devices change the antenna characteristics eliminating gain, thus transforming the device into a lossy antenna [2], [3], [8], [14], [16], [17]. These compact devices employ confined apertures to receive RF signals. Nevertheless, weak RF signals can cause the unavailability process to modulate the optical signal [18]. A successful modulation process determines the quality of the modulated signal resulting from EO conversion. Meanwhile, considering these circumstances, the integrated antenna-EO modulator is better designed as a Mach-Zehnder modulator (MZM); however, this design still delivers optical carrier-to-first sideband ratios (OCSRs) of approximately 49 dB [19].

The associate editor coordinating the review of this manuscript and approving it for publication was Zeev Zalevsky.
and 45 dB [20]. An integrated erbium-doped waveguide with an EO modulator and a pump power of 20 dBm can give a relative gain of 9.3 dB; nevertheless, this waveguide harmonic distortion and OCSR [21], meanwhile other design delivers 36 dB OCSRs [22]. An OCSR of 1 dB is necessary to improve the performance of an RoF link [23], [24].

A reflector can potentially optimize the RF input. A conventional paraboloid antenna can increase the RF output signal by up to 20 dB when focused on the patches [25], a lens reflector can provide an approximately 16 dBi gain [26], [27] and a plate lens can provide a 32 dBi gain [28]. A low profile reflector delivers a 48 dBi gain [29], 3D printed reflectors can provide an additional 22 dB to the RF input signal [30] and a Cassegrain antenna provides a 39 dBi gain [31]. The RF input signal can be increased from 16 to 48 dBi by using a reflector [19], [20], [22], [23], [24]. Overall, the use of a reflector can optimize the RF input signal in integrated antenna-EO modulator devices and RoF links by using an external EO modulator. However, external modulation in the external EO modulator is unlike the internal modulation in a directly modulated light source that can cause more chirp signals [32].

Nevertheless, the increased RF input signal can result in poor performance of the RoF link. These conditions can be caused by the nonlinear characteristics of the EO modulator [33] and, fiber optic link [34], in the form of harmonic distortions, in addition to single-mode fiber optic dispersion. Some distortions occur from phase noise [35] and thermal noise [36].

Harmonic distortion is a multiple of the RF frequency, $n f_{RF}$ [33], [37]. Unlimited power in the RF input can distort the RoF link and cause an increase in harmonics [38]. Harmonic distortion can be decreased by using certain linearity enhancement mechanisms of the EO modulator in the optical link or by reducing the effects of the EO modulator after the photodiode [37], [39]. Several techniques have been developed to decrease harmonics in the optical link by increasing the linearity of optical modulation.

Some of these techniques include the introduction of a dual-wavelength of the light source [39], [40], [41] in the RoF link and an improvement in the EO modulator by using parallel and serial MZMs [42], [43], [44], [45], [46], [47], [48]. An alternative method uses different polarizations of optical signals [49]. Previous technique add a wavelength division multiplexing (WDM) after the MZM and a coupler before MZM [50]. Other method uses a phase modulation before the signal enters the MZM input [48], [51]. However, these linear improvement require the installation of additional devices in the optical link [39], [40], [41], [42], [43], [44], [45], [46], [47], [48], [49], [50], [51]. These devices lead to intermodulation distortion and require an additional solution for either the optical link or the process after the photodiode. Intermodulation will appear if there are two or more tones at the transmitter [52], [53], [54]. Nonetheless, this condition can be diminished by minimizing the amount of equipment used in the optical link. Exploring the MZM transmission, sensitivity or active area [39], [41] is one solution to decrease harmonic distortion and intermodulation.

Adjusting the bias point voltage in the MZM transmission to increase the RF signal or reduce the noise figure is discussed in [55]. However, this mechanism requires additional devices to adjust the bias point. The analytical method of the MZM and EO silicon transfer function using the Taylor series [33] aims to explain a nonlinear model in harmonics and the change in its refractive index. The presence of the EDFA optical amplifier in the RoF link still gives less power to the main RF than the second harmonic, even at the initial RF input power [33]. However, this method still uses a phase shifter in one of the MZM arms and gives rise to intermodulation that needs to be addressed. Due to the dual-wavelength of a light source and the adjustment of the bias point near the minimum transmission point [41], this technique brings more power to the second harmonic than to the main RF at an RF input power of 8 dBm. Another method using linearity enhancement is MZM design [52]; however, the resulting MZM characteristic curve is quite different from the ideal curve, which is a cosine curve. Modeling an MZM distortion using a transfer function with phase change imbalances and fabricating a modulator using the Y branch to control the extinction ratio and chirp is discussed in [56]. This method requires four direct current (DC) biases to linearize the modulator. Expanding the Taylor series as a transfer function model is discussed in [57]; while this mechanism can minimize the second intermodulation (IMD2), and does not include the fiber optic dispersion.

Previous techniques [33], [41], [52], [56], [57] have shown that exploring or designing MZM transmission using the transfer function model can decrease harmonics distortions. Nevertheless, these techniques still need additional devices. Furthermore, there are restrictions on the scope of investigation of the MZM transmission area to confine the harmonics without adding more devices to the link. Specifically, to the best of our knowledge, no study has explored the allowable RF input amplitude through the MZM transmission boundary area, which is defined by the analytical and experimental investigation of MZM characteristics. This research is proposed to confine harmonics by maximizing the RF input power in the MZM transmission boundary area. An analytical investigation using a nonlinear model was proposed, investigated, and tested.

In detail, this study aims to find the allowable maximum RF input signal in the MZM transmission boundary area. The MZM transmission boundary area mechanism is convenient and can be used to confine harmonics in the RoF link without an optical amplifier, optical filter, or other optical devices. Therefore, we only need to explore the characteristics of MZM. The techniques described in this study involve minimal additional devices in the RoF link, a RoF link with simple structure, a lower optical power input, an optimized RF input amplitude to confine second harmonic, and no intermodulation. There is no intermodulation in this experiment because the maximum RF input amplitude in
the MZM transmission boundary area only uses a single tone.

In addition, the maximum allowable RF input amplitude is determined from the main RF-to-harmonics ratio (MHR). To obtain a comprehensive RoF performance, signal-to-noise-harmonic-ratio (SNHR) analysis and error vector magnitude (EVM) are performed in the analog RoF link. The results of this study show that this investigation gives a better EVM for QPSK modulation and matches the 3GPP (The 3rd Generation Partnership Project) Release 16.

II. NONLINEAR MODEL ON THE EO MODULATOR TRANSMISSION BOUNDARY AREA

A. NONLINEAR MODEL

We chose the MZM because it is a commonly used EO modulator at present. Fig. 1 shows the RF input power of the MZM, \( P(t) \) with a single carrier of the RF input, \( V_s(t) \) as shown in (1).

\[
V_s(t) = V_m \cos \omega_m t
\]

where \( V_m \) is RF input amplitude, and \( f_m \) is the RF frequency in the form of \( \omega_m \). The MZM DC bias input voltage, \( V_b \) will determine the electrical field generated by the MZM. The optical input power is \( P_{in} \), and the light source output signal is \( P_{LS} \), therefore \( P_{LS} = P_{in} = |E_{in}|^2 \), that is the square of the electric field input with the optical signal’s electric field input expressed as \( E_{in} = E_0 e^{j\omega_0 t} \). The electrical field generated at the electrodes between the MZM arms produces a change in the crystal refractive index of the optical waveguide, and the optical signal can modulate the RF signal from the input electrodes. It can be expressed as an electro-optical effect in the initial modulation process.

The MZM electrical field \( E_{MZM} \) in (2) [53] is found from the optical input power of the MZM, \( P_{in} \)

\[
E_{MZM} = \frac{1}{2} \left[ 1 + e^{j(\pi \Gamma_n \cos \omega RF t)} \right] \sqrt{2P_{in} e^{j\phi}},
\]

where \( V_n \) is either the MZM half-wave voltage or MZM switching voltage, \( \Gamma_m \) and \( \phi \) are the modulation index and the depth-of-intensity modulation respectively, and the phase shift constant can be written in (3) and (4).

\[
\Gamma_m = \frac{V_m}{V_n}
\]

\[
\phi = \frac{2\pi}{\Gamma_m} V_b
\]

Equations (3) and (4) are substituted into (2) and, we can reformulate (2) as (5).

\[
E_{MZM} (t) = \frac{1}{2} \left[ 1 + e^{j(\pi \Gamma_n \cos \omega RF t)} \right] \sqrt{2P_{in} e^{j\phi}},
\]

The electrical field generated by the MZM in (5) is the result of the interaction of \( V_s(t) \) with the refractive index of the optical material nonlinearity, which acts as the optical waveguide. This nonlinear condition can cause the appearance of RF harmonic signals that appear periodically. We can model the existence of this RF harmonic signal with (5), which is expanded by the Jacoby-Anger expansion as follows.

\[
e^{j\omega_0 t} = \sum_{n=-\infty}^{\infty} J_n (\omega_0) e^{j\phi}
\]

\[
e^{j\omega_0 t} = J_0 (\omega_0) + 2 \sum_{n=1}^{\infty} J_n (\omega_0) \cos (n\theta)
\]

We substitute (6) and (7) into (5) where \( n \) is the multiple of the harmonics that appear periodically in each multiple of the RF frequency therefore, we can model the electrical field equation output of the MZM in (8).

\[
E_{MZM} (t) = \frac{1}{2} \left[ 1 + e^{j\phi} (J_0 (\pi \Gamma_m) + \sum_{n=1}^{\infty} 2J_{2n} (\pi \Gamma_m) \cos (2n\omega RF t)) \right] \sqrt{2P_{in} e^{j\phi}}
\]

where \( J_{2n+1} (\pi \Gamma_m) \) and \( J_{2n} (\pi \Gamma_m) \) are the \( n \)th first kind of the Bessel function. Therefore, if we expand (8) again using (6) to obtain (9), as shown at the bottom of the next page, the electrical field output of the MZM consists of odd harmonics, as shown in (10), as shown at the bottom of the next page, and even harmonic as seen in (11), as shown at the bottom of the next page,.. Equation (9) explains the optical spectrum of the harmonics. The distribution of these harmonics is a function of the intensity modulation index as reflected by the parameter \( \Gamma_m \). Therefore, we model the MZM electrical field signal as simplified in (10) and (11), where \( n = 0 \) in \( E_{even} \), and we obtain (12) as follows:

\[
E_c (t) = e^{j\phi} e^{j\phi} \sqrt{2P_{in} J_0 (\pi \Gamma_m)}
\]

Equation (12) models an optical carrier signal. The nonlinear model in (9) is used to optimize the RF input power in the form of the main-RF-to-harmonics ratio (MHR). We obtain the RF output power using \( P_{RF} = |E_{MZM} E_{MZM}^{*}| \) to compare with the ITU-T standard.

B. MZM TRANSMISSION BOUNDARY AREA

A grasp is used to show the allowable RF input in the MZM transmission boundary area. This graph is the MZM transfer function to model the modulation signal. The MZM transfer function compares the MZM output signal with the input
signal, which can be expressed in the form of optical power normalization [59].

We determine the MZM transfer function based on the nonlinear model MZM signal as shown in (9). The MZM transfer function is represented as \( \frac{P_{\text{MZM}}}{P_{\text{in}}} \) with \( P_o = P_{\text{MZM}} = |E_{\text{MZM}}| \cdot E_{\text{MZM}}^* \) [33]; therefore, (9) can be transformed into (13), as shown at the bottom of the page, where the transfer function is represented in (14), as shown at the bottom of the page. If the transfer function has no RF input signal remaining, \( \Gamma_m = 0 \), and (14) becomes (15).

\[
T(V_b) = e^{2\omega t} \left( 1 + 4e^{j\theta} + 4e^{2j\theta} \right) \tag{15}
\]

The ideal MZM transfer function for transmission characteristics is shown in Fig. 2. The MZM has a voltage-transmission modulation area, \( V_{\pi} \). Note that, we use the term MZM boundary area to describe the MZM transmission area between the maximum and minimum transmission points. The black curve of the transfer function shows the transmission area of the MZM, where nearly linear area is approximately \( \frac{V_{\pi}}{2} \), where the MZM can work optimally as a modulator. Ideally, the most linear point is \( \frac{V_{\pi}}{2} \) with a transfer function at zero volts, and the normalized optical power is 0.5. The MZM transmission area is a periodic area of \( 2\pi \).

Fig. 2 shows that the MZM boundary area confines the optical transmission where the MZM can operate optimally with less harmonics distortion. It works optimally on the quadrature bias point (QBP) voltage, \( V_{\text{QBP}} \). QBP is a middle point of the switching voltage, and \( (V_b, V_m) \) and \( V_b \) can be \( = \frac{V_{\pi}}{2} \) or \( = \frac{V_{\pi}}{2} \). If QBP is precisely in the middle of the switching voltage, it lies on \( \left( \frac{V_{\pi}}{2}, 0.5 \right) \) or \( \left( \frac{V_{\pi}}{2}, 0.5 \right) \). The MZM transmission boundary area is explored in (16) to obtain \( V_m \) as in (16).

\[
V_m = \left( \frac{V_{\pi}}{2} \pm mV_{\pi} \right) \tag{16}
\]

The value of \( V_m \) is the RF input amplitude and has a maximum value of \( V_{\pi} \). If \( m \) is half of the peak-to-peak voltage \( V_{pp} \) and is determined by the boundary area, then \( 0 < m < 0.5 \). This study uses three values of \( m \): 0.2, 0.3, and 0.4. Nevertheless, \( V_m \) should be less than \( V_{\pi} \), to minimize the nonlinear distortion. Therefore, we assume that \( V_m \) is the RF

\[
E_{\text{MZM}}(t) = \frac{1}{2} \sqrt{2P_{\text{in}}} e^{2\omega t} \left( 1 + 2e^{j\theta} \left( \sum_{n=-\infty}^{\infty} J_{2n+1}(\pi \Gamma_m) e^{j(2n+1)\omega RF t} \right) + J_{2n}(\pi \Gamma_m) e^{j2n\omega RF t} \right) \tag{9}
\]

\[
E_{\text{odd}} = e^{j\theta} \sqrt{2P_{\text{in}}} \sum_{n=-\infty}^{\infty} J_{2n+1}(\pi \Gamma_m) e^{j(2n+1)\omega RF t} \tag{10}
\]

\[
E_{\text{even}} = e^{j\theta} \sqrt{2P_{\text{in}}} \sum_{n=-\infty}^{\infty} J_{2n}(\pi \Gamma_m) e^{j(2n+2)\omega RF t} \tag{11}
\]

\[
P_{\text{MZM}}(t) = P_{\text{in}} e^{2\omega t} \left[ 1 + 2e^{j\theta} \left( \sum_{n=-\infty}^{\infty} J_{2n+1}(\pi \Gamma_m) e^{j(2n+1)\omega RF t} \right) + J_{2n}(\pi \Gamma_m) e^{j2n\omega RF t} \right]^2 \tag{13}
\]

\[
T(V_b) = e^{2\omega t} \left[ 1 + 2e^{j\theta} \left( \sum_{n=-\infty}^{\infty} J_{2n+1}(\pi \Gamma_m) e^{j(2n+1)\omega RF t} \right) + J_{2n}(\pi \Gamma_m) e^{j2n\omega RF t} \right]^2 \tag{14}
\]
input amplitude and that the boundary areas for $V_m$ are 0.7$V_\pi$, 0.85$V_\pi$, and 0.9$V_\pi$. These three values of $m$ are investigated for the optimized RF input amplitude to limit harmonics. For these RF input amplitudes, we assume there is an RF optimized signal with tolerable harmonics and the absence of an optical filter, or optical amplifier in the RoF link.

Fig. (3) shows a detailed analytical investigation of the MZM transmission area with three points, the maximum bias point (MBP) at (0, 1), the minimum bias point (mBP) at (0, $V_\pi$), and the quadrature bias point (QBP) at ($V_\pi$, 0.5). A red solid line represents a line through the MZM transmission area at the QBP. A straight line confines the transmission boundary area through the QBP as a liner point in the MZM transmission area. The slope of the line through the QBP, $\frac{dT(V_b)}{dV_b}$ is shown in (17) if there is no RF input signal.

$$\frac{dT(V_b)}{dV_b} = 4j \frac{\pi}{V_\pi} e^{2\ln 4} \left(e^{j \frac{\pi}{\pi}} V_0 + 2e^{j \frac{2\pi}{\pi}} V_0\right)$$

In the beginning, we assume that this investigation in the MZM transmission boundary area confines the RF input values for the first, second, and third boundaries based on the RF input amplitudes of 0.7$V_\pi$, 0.85$V_\pi$, and 0.9$V_\pi$. The first boundary is shown by a rectangle with a dashed-dotted line with $V_m$ value of 0.7$V_\pi$. The solid line rectangle produces the second boundary as indicated by the blue arrow on the horizontal axis, which is 0.8$V_\pi$. The transmission boundary area with the dotted rectangle represents the third boundary with an RF input amplitude of 0.9$V_\pi$.

Furthermore, we examine the minimum of $V_m$ in the simulation; therefore, there will be no modulation process and no harmonics according to the RoF link simulation. This transmission boundary area is simulated using OptiSystem by determining the allowable RF input amplitudes $V_m$, based on the modeling and simulation results discussed in the next section.

C. FIBER IMPAIRMENT

Transmitting a signal with some performance criteria is influenced by fiber impairments that limit the fiber length in the RoF system. These fiber impairments are attenuation, dispersion, nonlinear distortions [39], and noise by random propagation effects such as polarization mode dispersion [59]. Attenuation occurs due to impurities material in the fiber optic and causes optical power degradation along the fiber length [60]. For this study, the electrical field output of the MZM in (8) is modeled in the frequency domain with the length of fiber, $z$ by using the Fast Fourier Transform $F_{\text{E}_{\text{MZM}}}(\omega, z = 0)$ and includes the fiber impairment in a fiber transfer function [32]:

$$\mathcal{H}(\omega, z) = 10^{-\alpha_{\text{loss}} z} \times e^{-j \beta_{\text{long}} z} \times e^{-j \beta_{\text{long}} (\omega - \omega_0) z} \times e^{-j \frac{\partial}{\partial t} |_{\omega_0} (\omega - \omega_0)^2 z}$$

where $\alpha_{\text{loss}}$ is the field attenuation coefficient ($\frac{\partial b}{\partial z}$), $j \beta_{\text{long}}$ is the propagation constant that represents the group delay per unit length, and $\frac{\partial b}{\partial t} |_{\omega_0} (\omega - \omega_0) = D \frac{\lambda}{\omega_0} \omega$ is the chromatic dispersion coefficient ($\frac{d^2 \beta}{d\lambda^2}$). For this study, the term $10^{-\alpha_{\text{loss}} z}$ is feasible to use because the confining harmonics of the MZM need some assumptions to simplify (18); therefore we can explore the nonlinear behavior in the MZM transmission boundary area.

D. ROF LINK SIMULATION

We select three RF input amplitudes and subsequently perform RoF simulation. We use several experimental data parameters in the simulation. The purpose of this simulation is to determine the value of $V_m$ in the transmission boundary area; that results in an acceptable number of harmonics after the photodiode output and an RF output power that fits the ITU-T standard. Furthermore, to determine the MZM output power, $P_{\text{MZM}}$ and the photodetector output, $P_{\text{RoF}}$, we add noise disturbances that predominantly appear in the RoF link, such as relative intensity noise (RIN) in light sources and thermal noise in photodiodes. RIN can be shown by (19) as discussed [61] where $k_{\text{RIN}}$ is 150 dB/Hz. For this simulation, 3 dB RIN noise is used and for the thermal noise photodetector in (19), the value of $10^{-0.022}$ W/Hz [62] is selected.

$$\sigma_{\text{RIN}}^2 = k_{\text{RIN}} I_{\text{dc}}^2 \Delta f$$

$$\sigma_{\text{thermal}}^2 = \frac{4k_b T}{R_L} \Delta f$$

Fig. 4 shows the RoF link configuration in the simulation, where the signal generator generates the RF signal. The optical input signal is a continuous wavelength (CW) laser with RIN noise. The end of the RoF link is a PIN photodiode output and an RF spectrum analyzer to explore the RF output with its harmonics. Meanwhile, we choose a PIN for the simulation and the experiment that justifies the MZM transmission boundary area condition.

The parameters for the CW laser, SD MZM, and signal generator are summarized in Table 1. The photodetector

| Component                | Parameter          | Value | Units |
|--------------------------|--------------------|-------|-------|
| SD-MZM (Lithium)         | Bias voltage 1 (QBP-1) | -4.5  | V     |
| Niobate with phase shift type | Bias voltage 2 (QBP-2) | -1.8  | V     |
|                          | Normalize electrical signal (uncheked) |       |       |
| Light source             | Wavelength         | 1550  | nm    |
| (Continuous-wave laser)  | Power              | -6.8  | dBm   |
|                          | Linewidth          | 10    | MHz   |
|                          | Noise dynamic      | 3     | dB    |
| Signal Generator (Sinewave) | Amplitude ($V_m$) | 0.6-1 | V     |
TABLE 2. Setting the parameters of the receiver.

| Component    | Parameter   | Value | Units |
|--------------|-------------|-------|-------|
| PIN Photodetector | Responsivity | 1     | A/W   |
|              | Dark current | 10    | nA    |
|              | Thermal noise | 10^{-6} | W/Hz |

TABLE 3. Setting parameters of the experiment.

| Components                      | Specifications                          |
|---------------------------------|----------------------------------------|
| Light source DFB                | Thorlabs’ 4-Channel, Fiber-Coupled      |
| Fiber optic from the light source to SD-MZM | Coa ring single-mode insensitive bending, ITU-T G.957 (Square connector) |
| Electro/optic modulator         | Single Drive-Mach Zehnder Modulator Lucent 2623NA |
| Signal Generator                | RIGOL DSG821                            |
| RF Signal Analyzer              | Anritsu MS2830A                         |
| Fiber optic from SD-MZM to OPM  | Single-mode insensitive bending, ITU-T G.957 |
| Optical connector               | Square connector with angle polished connector (SC-APC) |
| Photodetector                   | Ultrafast fiber optic photodetectors PIN UPD-15-IR2-FC |

FIGURE 4. RoF simulation.

parameters are listed in Table 2. The photodetector output, $P_{RF}$, is observed using a power meter and a dedicated RF spectrum analyzer to evaluate the RF signal harmonics. We add a fifth-order Chebyshev low-pass filter with a 0.5 dB ripple factor and a cutoff frequency of 2 GHz compared for the photodetector output signal. The Chebyshev LPF was chosen because we aim to obtain a sharp slope response that allows a ripple on the passband or stopband. We do not use BPF for this simulation because we observe the harmonics in the upper sideband.

III. RESULTS AND DISCUSSION

A. EXPERIMENTAL INVESTIGATION OF THE MZM TRANSMISSION BOUNDARY AREA

Two experiments are performed to justify the nonlinear model; the first is the MZM characteristics to obtain the MZM transmission boundary area, and the second is the RoF link to obtain the harmonics as illustrated in Fig. 5.

We use different bias voltage values on the MZM to obtain the MZM characteristics such as the QBP, MBP, and mBP as well as the MZM switching voltage. The MZM characteristics are expressed in terms of bias points and switching voltage values in the nonlinear model and analytical investigation parameters. Fig. 5 shows the experimental investigation. The first experiment is observed using an optical power meter at the MZM output. Meanwhile, the second is observed using an RF spectrum analyzer at the PIN output. These parameters are shown in Table 3.

We measure the MZM transmission boundary area using a single drive MZM of the Lucent 2623NA type connected to the optical pigtails on the input and output sections. These pigtails have square connectors (SCs) with a ferrule angle polished connector (APC) type. SD-MZM has a continuous wave (CW) optical input with a wavelength of 1550 nm connected using a single-mode insensitive bending optical fiber, ITU-T G.957. The connector from the SD-MZM input is connected via an adapter that connects the patch cord with the CW laser. A Lucent 10 Gb/s C-band MZM is used. The optical power signal is $P_{in}$, and the light source output signal is $P_{LS}$, as a result, $P_{LS} = P_{in} = |E_{in}|^2$ equals the square of the electric field input, with the optical signal electric field input represented as $E_{in} = E_0 e^{j2\pi fc t}$.

The other MZM input, DC bias voltage, $V_b$ is chosen because the MZM is a transverse modulator and it is measured in volts. In the first experiment, the MZM output power signal, $P_{MZM}$ is measured with an optical power meter (OPM) and the output power, $P_{MZM} = P_{o}$, is the value measured in the OPM.

The second experiment used a signal generator as the RF input and was observed at the PIN output. The results are investigated on the RF spectrum analyzer. The second experiment uses a continuous wave DFB laser that is operated at an output power of $-5.8$ dBm and wavelength of 1550 nm. The RF input is a sine wave with a frequency of 2 GHz. The MZM is biased at the quadrature point.

B. MZM OUTPUT POWER MODEL

Based on the measurement results, we obtain the MZM characteristic curve at QBP (1.5 V; 0.425 mW). Meanwhile, the model in (13) produces an ideal cosine-squared curve; therefore, if the initial phase model is based on [51], and from the measurement results, we can obtain a phase shift constant.
in (21) when there is no RF signal.

\[ \emptyset = \pi \left( \frac{V_b - V_\pi}{2} \right) \] (21)

Fig. 6 shows the measurement and modeling curves used to characterize the MZM transmission area. The measurement curve without RF input is represented by a solid black curve at a mBP of 0 V and QBP at 0.5 V; we obtained an MZM output power of 0.425 mW.

We substituted (21) into (14) and the nonlinear model is approximately equal to the measured MZM characteristic curve as shown by the solid blue curve. A solid black curve with squares is the MZM measurement curve without RF input. MZM has a switching voltage of 21 volts from the MBP at (−9 V, 0.85) to the lowest mBP at (12 V, 0) and when the QBP is at 0.5 V. A red dashed-dotted box outlines the MZM transmission boundary area. The analytical investigation is shown in Fig. 6.

C. CONFINE HARMONICS DISTORTION

1) OPTIMIZING THE RF INPUT MODEL

Consequently, we need to evaluate whether harmonic distortion will occur in the RF output signal. Fig. 7 shows the RF output power in the time domain (ns).

FIGURE 6. MZM characteristics model is based on measurement and modeling.

FIGURE 7. RF output power in the time domain (ns).

2) OPTIMIZING THE RF INPUT SIMULATION

We need to evaluate the domain effect of the main RF and harmonics in the simulation based on the RF input amplitude. Optimizing the RF input amplitude will increase the RF output power, nonetheless, when the total harmonics power is larger than the main RF output, the RF output power will decrease. We can see the domain effect of the main RF and harmonics in Fig. 8. The red gradient area within the RF input amplitude of 1.6V_\pi is the domain effect of the main RF. Meanwhile, harmonics become dominant after an RF input amplitude of 1.6V_\pi.

For this simulation, the RF output power is observed to obtain the power difference ratio that results from applying unfiltered and filtered signals after the photodiodes in the simulation. We simulate the unfiltered RF output power, P_{RF}, and compare the results to the filtered signal in Fig. 8. The solid red curve is the RF output power before the filter, which is larger than the filtered output signal, especially after the RF input amplitude reaches 2V_\pi volts. The solid blue curve with blue square markers is the RF output power after the LPF.

Based on the P_{RF} results before and after passing through the LPF, the black dashed-dotted curve shows the power ratio. This result is shown in Table 4. The increase in RF input amplitude will result in a greater power difference in RF output. However, for the RF input amplitude in the transmission boundary area at V_m \leq 0.9V_\pi, the power ratio is approximately 0%. In addition, the RF input amplitudes range from V_\pi to 5V_\pi, and the average power ratio is approximately 1.7%. When, V_m ranges from 5V_\pi to 10V_\pi, the power ratio difference is 8.5%. The most prominent effect of using a filter results in a power ratio of approximately 12% and the corresponding RF input amplitude is over 10V_\pi volts. Therefore, we choose 0.8V_\pi as the limiting RF input amplitude as it is the value before the harmonics distort the RF main signal, and it results in a power ratio 0%.

In Fig. 8, there is an RF input amplitude threshold, V_m threshold, which is the limiting value that can be used to design an integrated antenna-E0 modulator more than the
limiting value the modulation process still occurs. The RF input amplitude in the device must be greater than the RF input amplitude threshold. The detailed RF input amplitude threshold in the simulation is shown when the RF input amplitude is less than 0.003 $V_\pi$ or according to (16), when $m \leq 0.003$, in which there will not be any modulation signal. The minimum RF input threshold is larger than 0.003 $V_\pi$; therefore, the modulation process will occur. The device design requires a narrow MZM transmission area by using a low switching voltage. As discussed in [63], a low switching voltage or drive voltage can stabilize the modulation process using higher RF input power.

3) SECOND HARMONIC

Figure 9. (a) and (b) show the experiment as the validation of the simulation proof. Black solid curves show the main RF signal and the black dashed curves are the second harmonics in power. The main RF signal in the experiment gives a 17 dB difference from the second harmonic at an RF input of 0.8 $V_\pi$; otherwise, the simulation gives 15.5 dB. Nevertheless, the main RF of the experiment has an approximately 12.25 dB difference from the simulation.

Some transmission impairment, receiver noise floor [34], the total phase difference of the optical carrier, and the RF signal or phase mismatch in the link [51] cannot be included in the simulation. In [51], the researchers obtain 18 dB at the 90° phase from the main RF to each sideband harmonic.

4) HARMONICS TO THE MAIN RF RATIO

Based on nonlinear models, simulations, experiments, and analytical investigation within the MZM transmission boundary area, we can optimize the RF input amplitude until 0.9 $V_\pi$ is reached to confine second harmonics by using a filter after the PIN and 0.8 $V_\pi$ RF input amplitude without an optical filter. To examine the influence of thermal noise and harmonics, we included the PIN thermal noise in the nonlinear model and compared the RF output power to the ITU-T standard. We used the results to estimate the RF input optimization and confine the harmonics. We propose comparing the harmonics to the main RF power in the form of the total harmonic-power-to-main RF ratio (HMR) in percentage.

Based on (9), the value of $n$ ranges from $-10$ to 10 for the harmonics. Afterward, (9) will transform into (22), as shown at the bottom of the next page. Therefore, the HMR can be written as in (23) and (24), as shown at the bottom of the next page. In contrast, we obtain (25), as shown at the bottom of the next page, for the main RF-to-harmonics (MHR) ratio, and we use dB for this parameter.

$$10 \log_{10} (MHR) = \frac{P_{RF\ main}}{P_{harmonic}}$$ (26)

Based on (23), the results are illustrated in Figure 11(a) where the RF input amplitude determines the HMR value. Fig. 10 (a) shows the HMR with RF input amplitude chosen in the analytical investigation of the MZM transmission boundary area. The ratio reaches 100% at 2 $V_\pi$; and reaches under 10% at RF input amplitude within $V_\pi$ as detailed in Table 5. This comparison can be formulated as (25), the main-RF-to-harmonics ratio (MHR). Fig. 10 (b) shows that the weaker the RF input power is, the greater the MHR at the initial RF input amplitude.

The main RF is dominant until an RF input amplitude of 1.6 $V_\pi$; afterward, the total harmonics power is greater than the main RF. The possibility that the MHR can be infinite is determined by the RF input amplitude. Initially, the harmon-
ics are very low compared to the main RF. These low harmonics occur because the harmonics are approximately zero; therefore, the MHR peaks at 96 dB. Afterward, the MHR decreases with an increase in RF input amplitude because the harmonics begin to increase.

According to Table 5, if we want to optimize the RF input amplitude after the MHR peaks, the RF input amplitude must be below $V_\pi$; if we use amplitudes from $V_\pi$ to $2V_\pi$, the harmonics increase to until $1.6V_\pi$. The dominant effect of harmonics is after $1.6V_\pi$ and the main RF is no longer dominant. We can choose the RF input amplitude to confine the harmonics using the HMR.

5) ADJACENT CHANNEL LEAKAGE RATIO FOR NONLINEAR MODEL
The adjacent channel leakage power ratio (ACLR) is the ratio of the filtered mean power centered on the assigned channel frequency to the filtered mean power centered on an adjacent channel frequency at nominal channel spacing [64]. Fig. 11 shows the ACLR in different fiber lengths. We obtain ACLR by using (24) while including fiber attenuation. The blue solid, blue dashed, and blue dot curves are the transmission boundary areas. The blue dashed curve with a circle, and the star curve come from out of the transmission boundary area, and they occur at RF input powers of $V_\pi$ and $2V_\pi$, respectively.

3GPP Release 16 confines ACLR of the E-UTRA (Enhanced UMTS terrestrial radio access) at 30 dB. Fig. 11 shows that the nonlinear model can be used for 60 km at an RF input of 0.7, and 55 km at 0.8. For a single-carrier RF input amplitude in (1), without additional optical devices in the RoF link and while, considering only harmonics distortion and fiber attenuation, the nonlinear model matches 3 GPP Release 16.

6) SIGNAL-TO-NOISE RATIO AND SIGNAL TO NOISE, HARMONICS RATIO
The analog RoF performance is related to the signal-to-noise ratio, $\text{SNR}_{\text{RoF}}$. This $\text{SNR}_{\text{RoF}}$ is the photodiode output, which was compared with the noise in the RoF system. In the last step, we used the nonlinear model to estimate the signal-to-noise-harmonic ratio (SNHR) including thermal noise in the photodiode receiver. The noise power can be written as (27) [39].

$$P_{\text{optical}} = \sigma_{\text{thermal}}^2 \tag{27}$$

Based on HMR and MHR modeling in (25) and (26), respectively, we can estimate the SNHR by comparing the fundamentals of the RF output divided by the total harmonics added by quantum electronics noise in the receiver and, thermal

$$E_{\text{MZM}}(t) = \frac{1}{2}\sqrt{2P_{\text{in}}} e^{j\omega t} \left[ 1 + 2\sum_{0}^{\infty} (-jJ_{-5}(\pi \Gamma_m) e^{-j5\omega_{RF}t} + jJ_{-4}(\pi \Gamma_m) e^{-j4\omega_{RF}t} + (jJ_{-3}(\pi \Gamma_m) e^{-j3\omega_{RF}t} + J_{-2}(\pi \Gamma_m) e^{-j2\omega_{RF}t} + \cdots) \right]$$

$$HMR(\%) = \frac{P_{\text{harmonics}}}{P_{\text{mainRF}}} \times 100\% \tag{23}$$

$$HMR = \frac{\left( \left( \sum_{n=-\infty}^{\infty} J_{2n+1}(\pi \Gamma_m) e^{j(2n+1)\omega_{RF}t} \right)^2 + \left( \sum_{n=-\infty}^{\infty} J_{2n}(\pi \Gamma_m) e^{2n\omega_{RF}t} \right)^2 \right) - \left( J_1(\pi \Gamma_m) e^{\omega_{RF}t} \right)^2 - \left( -jJ_{-1}(\pi \Gamma_m) e^{-\omega_{RF}t} \right)^2 - J_0(\pi \Gamma_m) \right)}{(jJ_1(\pi \Gamma_m) e^{\omega_{RF}t} \right)^2 + \left( -jJ_{-1}(\pi \Gamma_m) e^{-\omega_{RF}t} \right)^2} \tag{24}$$

$$A = \left( \sum_{n=-\infty}^{\infty} J_{2n+1}(\pi \Gamma_m) e^{j(2n+1)\omega_{RF}t} \right)^2$$

$$B = \left( \sum_{n=-\infty}^{\infty} J_{2n}(\pi \Gamma_m) e^{2n\omega_{RF}t} \right)^2$$

$$C = (jJ_1(\pi \Gamma_m) e^{\omega_{RF}t} \right)^2$$

$$D = \left( -jJ_{-1}(\pi \Gamma_m) e^{-\omega_{RF}t} \right)^2$$

$$E = J_0(\pi \Gamma_m)$$

$$HMR(\text{dB}) = \frac{A + B - C - D - E}{C + D} \tag{25}$$
TABLE 5. Dominant RF main signal and harmonics comparison.

| RF Input Amplitude (volt) | HMR (%) | MHR (dB) |
|--------------------------|---------|----------|
| 0.7Vπ                    | 4.5     | 13.617   |
| 0.8Vπ                    | 5.8     | 12.355   |
| 0.9Vπ                    | 7.5     | 11.139   |
| Vπ                       | 9.5     | 10.792   |
| 1.6Vπ                    | 27.5    | 5.740    |
| 2Vπ                      | 75      | 1.761    |
| 2.2Vπ                    | 100     | 0        |

7) ERROR VECTOR MAGNITUDE
We use modulation signals to check the efficacy of the nonlinear model. Fig. 13 shows the EVM (%) for some modulation signals in fiber length. There are three types of modulation signals, QPSK, 256 QAM, and PAM 4. QPSK curves are red, 256 QAM are blue, and PAM 4 are black. Each modulation signal is investigated for some RF input. The terms 0.7Vπ, 0.8Vπ, and 0.9Vπ are within the transmission boundary area. The 0.7Vπ curves are solid lines, 0.8Vπ curves are dashed, and 0.9Vπ are dashed-dotted curves, Vπ curves are dotted, and 1.2Vπ are dashed and double dotted. The figure shows that QPSK has a lower EVM than 256 QAM or PAM 4. The EVM (%) of QPSK matches the 3GPP specifications of Release 16. The limit is 17.5% for QPSK and the QPSK can give convenience for 45 km of fiber length. This fiber length is convenient for medium or small range of RoF link during distributed antenna system (DAS) implementation.

Afterward, (28) can be written as (29), and (30), as shown at the bottom of the next page. Fig. 12 shows the analog RoF performance in the form of SNHR and SNR. The blue solid curve shows the SNHR, while the dashed curve shows the SNR. Compared with SNHR in the MZM transmission boundary area, SNR has more power of 28 dB. SNR is shown by the red arrow in the figure. It can be assumed that the harmonics in the form of SNR are a primary source of RF power. In the form of SNHR, the harmonics are some distortion signals.

D. THIRD INTERMODULATION DISTORTION
This section discusses the third intermodulation (3rd IMD) in the nonlinear model and is shown in Fig. 15. The third intermodulation is the most dominant effect on the RF signal [65]. The intermodulation distortion occurs in dual-tone RF frequencies, f1 and f2. The third IMD occurs as (2f1 + f2), (2f1 − f2), (f1 + 2f2), and (−f1 + 2f2) [66].

In this study, for the nonlinear model, SD-MZM uses two RF frequencies: 2 GHz and 3.5 GHz. The RF signal equation is shown in (31).

\[
V(t) = V_{m1}\cos\omega_{m1}t + V_{m2}\cos\omega_{m2}t
\]  

(31)

Then (9) becomes (32).

\[
E_{MZM}(t) = E_c + (E_{odd\_RF1} + E_{even\_RF2} + E_{odd\_RF2} + E_{even\_RF2} + E_{IMD3})
\]  

(32)

The 3rd IMD is shown in (33), as shown at the bottom of the next page.

Fig. 15 shows the three curves, the blue solid curve is the main RF output, the red dashed curve is 2nd H, and the black dashed-dotted curve is 3rd IMD. The 2nd H power is lower than the main RF within 0.8Vπ of the RF input amplitude.

The 0.8Vπ is the limit of maximum RF input with tolerable 2nd H and 3rd IMD. For 0.9Vπ of RF input amplitude, the main...
RF output power becomes lower and decreases and then the RF input amplitude can only be optimized within a modulation index of $0.8\pi$. This nonlinear model is appropriate for RoF link without any additional devices and gives 78 dB for better main RF than 3rd IMD, despite the 2nd H being less than 4 dB from main RF.

The 3rd IMD reaches an output power of $-154$ dB at an RF input amplitude of 12.25 dBm. This condition is better than [50] due to the RF output power difference of 70 dB from the 3rd IMD. The techniques applied in [50] and [65] gave the main RF output more power of 50 dB difference from 3rd IMD. Meanwhile, in [67], the previous technique could give 2nd IMD suppression, but the 2nd H remained constant. The 3rd IMD in this study gives a better fundamental-RF-to-third-intermodulation ratio (FIR) of 20 dB than [65].

**IV. PROPOSED NONLINEAR AND OTHER NONLINEAR MODEL**

**A. OPTICAL CARRIER-TO-FIRST SIDEBAND RATIOS**

We evaluate the optical output signal to determine whether there is a distortion or larger harmonic power than the main RF signal. Fig. 16 (a), (b), and (c) show the optical signal and harmonics; generally, the higher the RF input power is, the more harmonics will occur. The first-order harmonic ($f_c + f_{RF}$) is optical with the main RF signal, which has the highest optical power, as shown in Fig. 16 (a). The second order is the optical signal with second harmonics ($f_c + 2f_{RF}$) which increases more than the third ($f_c + 3f_{RF}$), fourth ($f_c + 4f_{RF}$), and fifth ($f_c + 5f_{RF}$) harmonics. Fig. 16 (a) shows that the dominant harmonics are second and third harmonics and the

$$SNHR_{RoF} = \frac{MHR \cdot P_{\text{harmonic}}}{P_{\text{harmonic}} + P_{\text{optical}}}$$

$$SNHR_{RoF} = \frac{(E_1)^2 + (E_{-1})^2}{(E_{\text{odd}})^2 + (E_{\text{even}})^2 - (E_1)^2 - (E_{-1})^2 + \sigma_{\text{thermal}}^2}$$

$$E_{\text{IMD}3} = e^{j\theta} \sqrt{2P_{in}} \left( (-1) J_2 (\Gamma_{m1}) J_1 (\Gamma_{m2}) e^{j(2\omega_1 + \omega_2)t} + J_2 (\Gamma_{m1}) J_{-1} (\Gamma_{m2}) e^{j(2\omega_1 - \omega_2)t} + (-1) J_1 (\Gamma_{m1}) J_2 (\Gamma_{m2}) e^{j(\omega_1 + 2\omega_2)t} + J_{-1} (\Gamma_{m1}) J_2 (\Gamma_{m2}) e^{j(-\omega_1 + 2\omega_2)t} \right)$$
others are small in power. The second harmonic increases by approximately 73% compared to the main RF signal when the RF input amplitude is 0.8Vπ, as shown in Fig. 16 (b). Meanwhile, the second harmonic increases by 89% in power when the RF input voltage is 0.9Vπ, as shown in Fig. 16 (c).

Fig. 16 (a), (b), and (c) indicate that the second harmonic is the most dominant harmonic because its power distorts the main RF signal, whereas the other harmonics are still very weak and tolerable. Table 6 shows the detailed power in percentage from the second, third, fourth, and fifth harmonics in the optical signal. While the harmonics increase as the RF input amplitude increases, the first order decreases by 89% at an RF input amplitude of 0.8Vπ and 81% at 0.9Vπ. The optical power of the second harmonic is 83% compared to that of the first order, the main RF signal. These conditions show that if the increase in RF input amplitude is greater than 0.9Vπ, this condition causes the second harmonic to be greater than the main RF in the optical signal.

From the RF input amplitude based on the analytical investigation in the MZM transmission boundary area, we obtain OCSRs of 2.91 dB for 0.7Vπ, 2.365 dB for 0.8Vπ, and 1.97 dB for 0.9Vπ. This OCSR is better than the OCSRs in [19] and [20]; nevertheless, the best OCSR is 1 dB.

B. RF OUTPUT POWER

A previous technique [36] used an EDFA optical amplifier in the RoF link; and gave the main RF power less than the second harmonic at a wavelength of 1552.4 nm, even at the initial RF input power. Moreover, the previous technique did not include fiber dispersion in the link. The transmission boundary area uses no optical amplifier, optical filter, or any other optical devices except a CW laser, SD MZM, and PIN photodiode. Our technique gives better main RF power than the second harmonic from up to 25 dBm of RF input amplitude at a wavelength of 1550 nm. The main RF power and 2ndH are −83 dBm and −113 dBm, respectively; the main RF remains better than the 2ndH by approximately 30 dB, while [36], the main RF only remained approximately 20 dB from the 2ndH at a wavelength of 1552.1 nm. Our technique gives better main RF output power than the second harmonic from up to 12.46 dBm of RF input amplitude at a wavelength of 1550 nm. In this study, the main RF power and 2ndH are −76 dBm and −78 dBm, the main RF remains better than the 2ndH by approximately 2 dB in the transmission boundary area, while in [33], the main RF only remained lower by approximately 20 dB from the 2ndH at a wavelength of 1552.4 nm. Moreover, in [41], this technique gave the second harmonic more power than the main RF at an RF input power of 10 dBm due to the increase in the 2ndH.

V. CONCLUSION

We proposed an analytical investigation and experiment of the MZM transmission boundary area using nonlinear models. This technique is an alternative solution to improve the RF input in an integrated antenna-EO modulator, by using the maximum allowable RF input amplitude and by limiting the harmonics. We experimentally measured the MZM transmission area and theoretically investigated the allowable RF input on an analog radio over fiber link. We proposed a nonlinear model to obtain the RF output signal and its harmonics.
while varying the RF input amplitude. This model provide appropriate experimental MZM characteristics and could precisely evaluate the RF output with its harmonics. The model and simulated evaluated showed that within the MZM transmission boundary area, when the RF input was 0.8 times the MZM switching voltage, the RF output signal was not distorted; therefore, it was chosen as the boundary area to confine the harmonics without any additional optical devices. At this value, we obtained an HMR of 5.8%. We showed that the RF input amplitude was approximately 1.6 times the switching voltage within the MZM transmission area, and the main RF input was still dominant. We could optimize the RF input at any voltage up to its switching voltage, and obtain the HMR below 9.5%; however, the signal was distorted. After considering the optical fiber impairment, thermal noise, and harmonics distortion in the nonlinear model with modulation signal of a QPSK, the ACLR and EVMs could satisfy the 3GPP specification. After evaluating the EVMs for various small- or medium-sized ranges of RoF links.

ACKNOWLEDGMENT

The authors would like to thank Dr. Bambang Widiyatmoko and Dwi Hanto, Ph.D. of the Research Center for Photonics, National Research and Innovation Agency at Republic of Indonesia for assembling the experiment.

REFERENCES

[1] D. F. Paredes Páliz, G. Royo, F. Aznar, C. Aldea, and S. Celfa, “Radio over fiber: An alternative broadband network technology for IoT,” Jornada de Jóvenes Investigadores del I3A, vol. 8, pp. 1–8, Dec. 2020.

[2] N. Kohn, H. Murata, and Y. Okamura, “Electro-optic modulator using an antenna-coupled-electrode array and a polarization-reversed structure for a radar tracking system J-y,” Radio Sci. Bull., vol. 349, no. 349, pp. 32–39, 2014.

[3] Y. N. Wijayanto, A. Kanno, S. Nakajima, P. Daud, and S. Celfa, “Radio over fiber: An alternative broadband network technology for IoT,” Jornada de Jóvenes Investigadores del I3A, vol. 8, pp. 1–8, Dec. 2020.

[4] R. Degl’Innocenti, D. S. Jessop, Y. D. Shah, J. Sibik, J. A. Zeitter, P. R. Kidambi, S. Hofmann, H. E. Beere, and D. A. Ritchie, “Tera- hertz optical modulator based on metamaterial split-ring resonators and graphene,” Opt. Eng., vol. 53, no. 5, May 2014, Art. no. 057108.

[5] L. Novotny and N. van Hulst, “Antennas for light,” Nature Photon., vol. 5, no. 2, pp. 83–92, Feb. 2011.

[6] A. Melikyan, L. Allisotti, A. Muslija, D. Hillekuss, P. C. Schindler, J. Li, R. Palmer, D. Korn, S. Muehlbrandt, D. Van Thourhout, B. Chen, R. Dimu, M. Sommer, C. Koos, M. Kohl, W. Freude, and J. Leuthold, “High-speed plasmonic phase modulators,” Nature Photon., vol. 8, no. 3, pp. 229–233, Mar. 2014.

[7] M. Hochberg, T. Baeher-Jones, G. Wang, M. Shearn, K. Harwood, J. Luo, B. Chen, Z. Shi, R. Lawson, P. Sullivan, A. K.-Y. Jen, L. Dalton, and A. Scherer, “Tera- hertz all-optical modulation in a silicon–polymer hybrid system,” Nature Mater., vol. 5, no. 9, pp. 703–709, Sep. 2006.

[8] X. Zhang, “Integrated photonic electromagnetic field sensor based on broadband bowtie antenna coupled silicon organic hybrid modulator,” J. Lightw. Technol., vol. 32, no. 20, pp. 3774–3784, Oct. 1, 2014.

[9] M. S. Eggleston, K. Messer, L. Zhang, E. Yablonovitch, and M. C. Wu, “Optical antenna enhanced spontaneous emission,” Proc. Nat. Acad. Sci. USA, vol. 112, no. 6, pp. 1704–1709, 2015.

[10] R. B. Waterhouse and D. Novak, “Integrated antenna/electro-optic modulator for RF photonic front-ends,” in IEEE MTT-S Int. Microw. Symp. Dig., Jun. 2011, pp. 9–12.

[11] U. Ozkaya and L. Seyfi, “Dimension optimization of microstrip patch antenna in X/Ku band via artificial neural network,” Proc.-Social Behav. Sci., vol. 195, pp. 2520–2526, Jul. 2015.

[12] X. Zhang, “Antenna-coupled silicon-organic hybrid integrated photonic crystal modulator for broadband electromagnetic wave detection,” Photon. West, vol. 9362, Mar. 2015, Art. no. 93620Q.

[13] F. Qiu, A. M. Spring, D. Maca, M.-A. Ozawa, K. Odoi, I. Aoki, A. Otomo, and S. Yokoyama, “TiO2 ring-resonator-based EO polymer modulator,” Opt. Exp., vol. 22, no. 12, p. 14101, 2014.

[14] Y. Zhang, H. Wu, D. Zhu, and S. Pan, “An optically controlled phased array antenna based on single sideband polarization modulation,” Opt. Exp., vol. 22, no. 4, p. 3761, 2014.

[15] A. Afridi and E. Kocaba, “Beam steering and impedance matching of plasmonic horn nanoelements,” Opt. Exp., vol. 24, no. 22, p. 25647, 2016.

[16] D. H. Park, V. R. Pagán, T. E. Murphy, J. Luo, A. K.-Y. Jen, and W. N. Herman, “Free space millimeter wave-coupled electro-optic high speed nonlinear polymer phase modulator with in-plane slotted patch antennas,” Opt. Exp., vol. 23, no. 7, p. 9464, 2015.

[17] Y. Natali, P. S. Priambodo, and E. T. Rahardjo, “Radio frequency to lightwave signal using integrated antenna and optical material for electro optic alteration,” in Proc. 4th Int. Conf. Sci. Technol. (ICST), Aug. 2018, pp. 1–5.

[18] A. Chen and E. J. Murphy, Broadband Optical Modulators. New York, NY: CRC Press, 2012.

[19] Y. Matsuoka, T. Inoue, H. Murata, and A. Sanada, “Millimeter-wave band optical single-sideband modulator using array-antenna-electrode and polarization-reversed structures with asymmetric Mach–Zehnder waveguide,” Jpn. J. Appl. Phys., vol. 4, pp. 1–5, Jul. 2018.

[20] H. Murata, “Millimeter-wave-band electro-optic modulators using antenna-coupled electrodes for microwave photonic applications,” J. Lightw. Technol., vol. 38, no. 19, pp. 5485–5491, Oct. 1, 2020.

[21] M. Zhang, G. Hu, and X. Wang, “On-chip electro-optic modulator with loss compensation based on polymeric active-integrated waveguides,” IEEE Access, vol. 8, pp. 116470–116477, 2020.

[22] O. D. Herrera, “Silica/electro-optic polymer optical modulator with integrated antenna for microwave receiving,” J. Lightw. Technol., vol. 32, no. 20, pp. 3861–3867, Oct. 15, 2014.

[23] C. Lim, K.-L. Lee, A. Nirmalathas, D. Novak, and R. Waterhouse, “Technique to improve carrier-to-interference ratio of optical single sideband with carrier modulated signals,” Opt. Exp., vol. 14, no. 23, p. 11077, 2006.

[24] C. Lim, M. Attigalle, A. Nirmalathas, D. Novak, and R. Waterhouse, “Analysis of optical carrier-to-sideband ratio for improving transmission performance in fiber-radio links,” IEEE Trans. Microw. Theory Techn., vol. 54, no. 5, pp. 2181–2187, May 2006.

[25] Y. Natali, P. S. Priambodo, and E. T. Rahardjo, “Increasing received modulation signal for integrated EOM microwave antenna by using parabolic reflector,” in Proc. IEEE Conf. Antenna Meas. Appl. (CAMA), Oct. 2019, pp. 214–217.

[26] C. C. Cruz, J. R. Costa, C. A. Fernandes, and S. A. Matos, “Focal-plane multibeam dual-band dielectric lens for Ka-band,” IEEE Antennas Wireless Propag. Lett., vol. 16, pp. 432–436, 2017.

[27] J.-W. Lian, Y.-L. Ban, Z. Chen, B. Fu, and C. Xiao, “SIW folded Cassegrain lens for millimeter-wave multibeam application,” IEEE Antennas Wireless Propag. Lett., vol. 17, no. 4, pp. 583–586, Apr. 2018.

[28] M. R. D. Kodnoeih, Y. Letestu, R. Saulaue, E. M. Cruz, and A. Doll, “Compact folded Fresnel zone plate lens antenna for mm-wave communications,” IEEE Antennas Wireless Propag. Lett., vol. 17, no. 5, pp. 873–876, May 2018.

[29] V. Manohar, J. M. Kovitz, and Y. Rahmat-Samii, “Synthesis and analysis of low profile, metal-only stepped parabolic reflector antenna,” IEEE Trans. Antennas Propag., vol. 66, no. 6, pp. 2788–2797, Apr. 2018.

[30] A. Rebollo, F. Vaquero, M. Arrebola, and M. R. Pino, “3D-printed dual-reflector antenna with self-supported dielectric subreflector,” IEEE Access, vol. 8, pp. 209091–209100, 2020.

[31] H. R. D. Filgueiras, R. M. Borges, M. C. Melo, T. H. Brandao, and A. C. Souza, “Dual-band wireless fronthaul using a FSS-based focal-point Cassegrain antenna assisted by an optical midhaul,” IEEE Access, vol. 7, pp. 112578–112587, 2019.
A. M. Gutierrez, A. Brimont, J. Herrera, M. Aamer, D. J. Thomson, V. Gardes, G. T. Reed, J.-M. Fedeli, and P. Sanchis, “Analytical model for calculating the nonlinear distortion in silicon-based electrooptic Mach–Zehnder modulators,” J. Lightw. Technol., vol. 31, no. 23, pp. 3603–3613, Dec. 1, 2013.

G. Meslener, “Chromatic dispersion induced distortion of modulated monochromatic light employing direct detection,” IEEE J. Quantum Electron., vol. JQE-20, no. 10, pp. 1208–1216, Oct. 1984.

E. P. Martin, T. Shao, V. Vujicec, P. M. Anandarajah, C. Browning, R. Llorente, and Liam P. Barry, “25-GHz OFDM 60-GHz radio over fiber system based on a gain switched laser,” J. Lightw. Technol., vol. 38, no. 15, pp. 1635–1643, Apr. 15, 2015.

I. S. Amiri, F. M. A. M. Houssien, A. N. Z. Rashed, and A. E.-N. A. Mohammed, “Temperature effects on characteristics and performance of near-infrared wide bandwidth for different avalanche photodiodes structures,” Results Phys., vol. 14, Sep. 2019, Art. no. 102399.

V. A. Thomas, M. El-Hajjar, and L. Hanzo, “Millimeter-wave radio over fiber optical upconversion techniques relying on link linearity,” IEEE Commun. Surveys Tuts., vol. 18, no. 1, pp. 59–123, 1st Quart., 2016.

W. S. C. Chang, RF Photonic Technology in Optical Fiber Links. Cambridge, U.K.: Cambridge Univ. Press, 2002.

V. A. Thomas, M. El-Hajjar, and L. Hanzo, “Performance improvement and cost reduction techniques for radio over fiber communications,” IEEE Commun. Surveys Tuts., vol. 17, no. 2, pp. 627–670, 2nd Quartary, 2015.

S. Yu, T. Jiang, J. Li, R. Zhang, G. Wu, and W. Gu, “Linearized frequency doubling for microwave photonics links using integrated parallel Mach–Zehnder modulator,” IEEE Photon. J., vol. 5, no. 4, Jul. 2013, Art. no. 5501108.

R. Zhu, M. Hui, D. Shen, and X. Zhang, “Mach–Zehnder modulator modulated radio-over-fiber transmission system using dual wavelength linearization.” Opt. Commun., vol. 385, pp. 229–237, Feb. 2017.

H. Yamazaki, H. Takahashi, T. Goh, Y. Hashizume, T. Yamada, S. Minou, H. Kawakami, and Y. Miyamoto, “Optical modulator with a near-linear field response,” J. Lightw. Technol., vol. 34, no. 16, pp. 3796–3802, Aug. 15, 2016.

H. Zhang, L. Cai, S. Xie, K. Zhang, X. Wu, and Z. Dong, “A novel radio-over-fiber system based on carrier suppressed frequency eightfold millimeter wave generation,” IEEE Photon. J., vol. 9, no. 5, pp. 1–6, Oct. 2017.

M. Du, P. Zheng, J. Li, P. Liu, X. Xu, G. Hu, B. Yun, and Y. Cui, “Photonic generation of frequency-doubled microwave photonics links using cascaded Mach–Zehnder modulators,” Optik, vol. 240, Aug. 2021, Art. no. 166933.

Y. Zhou, L. Zhou, M. Wang, and Y. Xia, “Linearity characterization of a dual–parallel silicon Mach–Zehnder modulator linearity characterization of a dual–parallel silicon Mach–Zehnder modulator,” IEEE Photon. J., vol. 8, no. 6, pp. 1–8, Oct. 2016.

Y. He, Y. Li, Y. Cai, X. Zhang, J. Liu, J. Xiao, S. Chen, and D. Fan, “A full-duplex 100-GHz radio-over-fiber communication system based on frequency quadrupling,” Optik, vol. 175, pp. 148–153, Dec. 2018.

W. Li, W. T. Wang, W. Li, W. T. Wang, and N. H. Zhu, “Photonic generation of radio-frequency waveforms based on dual-parallel Mach–Zehnder modulator,” IEEE Photon. J., vol. 6, no. 3, pp. 1–8, 2014.

M. Noweir, “Digitally linearized radio-over fiber transmitter architecture for cloud radio access network’s downlink,” IEEE Trans. Microw. Theory Techn., vol. 66, no. 7, pp. 3564–3574, Jul. 2018.

B. Wu, J. Y. Sung, J. H. Yan, M. Xu, J. Wang, F. Yan, S. Jian, and G.-K. Chang, “Polarization-insensitive remote access unit for radio-over-fiber mobile fronthaul system by reusing polarization orthogonal light waves,” IEEE Photon. J., vol. 8, no. 1, pp. 1–8, Dec. 2016.

E. I. Ackerman and C. H. Cox, “Improved RF interference suppression method,” J. Lightw. Technol., vol. 38, no. 19, pp. 5546–5550, Oct. 1, 2020.

H. N. Parajuli and E. Udvary, “Novel vestigial sideband modulation scheme to enhance the SNR in radio over fiber systems,” Radioengineering, vol. 26, no. 4, pp. 961–967, Dec. 2017.

S. Jin, L. Xu, T. Manzur, and Y. Li, “Quantum well Mach–Zehnder intensity modulator with enhanced linearity for direct detection,” J. Lightwave Technol., vol. 35, no. 17, pp. 3785–3790, 2017.

P. S. Devgan, D. P. Brown, and R. L. Nelson, “RF performance of single sideband modulation versus dual sideband modulation in a photonic link,” J. Lightw. Technol., vol. 33, no. 9, pp. 1888–1895, May 1, 2015.

F. Xavier, Radio Over Fiber for Wireless Communication. Hoboken, NJ, USA: Wiley, 2014.

S. Shi, J. Yuan, Q. Huang, C. Shi, X. Luo, S. Lu, P. Yuan, H. Yu, and Q. Yue, “Bias controller of Mach–Zehnder modulator for electro-optic analog-to-digital converter,” Micromachines, vol. 10, no. 12, pp. 1–10, 2019.

Y. Yamaguchi, A. Kanno, T. Kawanishi, M. Isitsu, and H. Nakajima, “Precise optical modulation using extinction-ratio and chirp tunable single-drive Mach–Zehnder modulator,” J. Lightw. Technol., vol. 35, no. 21, pp. 4781–4788, Nov. 1, 2017.

N. J. Frigo, M. N. Hutchinson, and C. R. S. Williams, “Modelling non-linearities in RF photonic links,” J. Lightw. Technol., vol. 36, no. 19, pp. 4371–4378, Oct. 15, 2018.

P. Horvath and I. Frigyes, “Effects of the nonlinearity of a Mach–Zehnder modulator on OFDM radio-over-fiber transmission,” IEEE Commun. Lett., vol. 9, no. 10, pp. 921–923, Oct. 2005.

M. A. Saleh and C. T. Baha, Fundamentals of Photonics. Hoboken, NJ, USA: Wiley, 1994.

G. P. Agrawal, Fiber-Optic Communications Systems, vol. 6, 3rd ed. New York, NY, USA: Wiley, 2002.

M. U. Hadi, H. Jung, S. Ghaffar, P. A. Traverso, and G. Tartarini, “Optimized digital radio over fiber system for medium range communication,” Opt. Commun., vol. 443, pp. 177–185, Jul. 2019.

Z. Ibrahim, C. B. M. Rashidi, S. A. Aljunid, A. K. Rahman, and S. Yaakob, “PIN and APD analysis for optical CDMA based on the proposed radio over fiber (RoF) approach,” Int. J. Appl. Eng. Res., vol. 12, no. 7, pp. 1396–1400, 2017.

D. Borlaug, P. T. S. DeVore, A. Rostami, O. Boyraz, and B. Jalali, “Demonstration of Vπ reduction in electro-optic modulators using modulation instability,” IEEE Photon. J., vol. 6, no. 5, Oct. 2014, Art. no. 6802409.

LTE: Evolved Universal Terrestrial Radio Access (E-UTRA): User Equipment (UE) Radio Transmission and Reception (3GPP TS 36.101 Version 12.5.0 Release 12), document ETSI 3GPP TS 136 101-12.5.0 2020.

Z. Zhu, S. Zhao, X. Li, K. Qu, T. Lin, and B. Lin, “Dynamic range improvement for an analog photonic link using an integrated electro-optic dual-polarization modulator,” IEEE Photon. J., vol. 8, no. 2, pp. 1–10, Apr. 2016.

F. Vacondio, M. Mirshafiei, J. Basak, A. Liu, L. Liao, M. Paniccia, and L. A. Rusch, “A silicon modulator enabling RF over fiber for 802.11 OFDM signals,” IEEE J. Sel. Topics Quantum Electron., vol. 16, no. 1, pp. 141–148, 2010.

B. M. Haas and T. E. Murphy, “Linearized downconverting microwave photonic link using dual-wavelength phase modulation and optical filtering,” IEEE J. Photon., vol. 3, no. 1, pp. 1–12, Feb. 2011.
PURNOMO SIDI PRIAMBODO (Member, IEEE) received the B.S. degree in electrical engineering from Universitas Gadjah Mada, Indonesia, in 1987, the M.S. degree from the School of Electrical and Computer Engineering, Oklahoma State University, Stillwater, in 1996, and the Ph.D. degree in photonics and diffractive optics from the Department of Electrical Engineering, University of Texas at Arlington, in 2003. From 1987 to 1994, he joined PT Indosat. His job work experience spans from satellite earth stations to data communication and system programs. From 2003 to 2005, he was a Postdoctoral Fellow at UT Arlington and a Senior Research Scientist at Resonant Sensors Inc. He joined the Department of Electrical Engineering, Universitas Indonesia, in 2007, where he has been a position as an associate professor in optoelectronics engineering, since 2017. He has a U.S. patent disclosure on nonlinear optical guided-mode resonance filters and has published several papers on guided-mode resonance filters and devices. He has been a reviewer for various professional journals. He has been a member of the IEEE Lasers and Electro-Optics Society (LEOS), since 1996. He is also a member of the Honor Societies of Eta Kappa Nu and Phi Beta Delta. His research interests include diffractive optics, optical interconnection, quantum optoelectronics, semiconductor lasers, finite-difference time-domain electromagnetic propagation analysis, modulator/receivers, optical waveguides, photonic crystals, quantum electrodynamics, and free-space optical communications.

EKO TJIPTO RAHARDJO (Member, IEEE) was born in Pati, Indonesia, in April 1958. He received the I.r. (Insinyur) degree from Universitas Indonesia, Depok, Indonesia, in 1981, the M.S. degree from the University of Hawai‘i at Manoa, Honolulu, HI, USA, in 1987, and the Ph.D. degree from Saitama University, Urawa, Japan, in 1996, all in electrical engineering. Since 1982, he has been a Teaching Assistant with the Department of Electrical Engineering, Universitas Indonesia. Since 2005, he has also been appointed as a Professor in electrical engineering. He was the Chairperson of the University Senate, Universitas Indonesia, from 2011 to 2012; the Head of the Department of Electrical Engineering, Universitas Indonesia, from 2004 to 2008; and the Executive Director of the Quality Undergraduate Education (QUE), Department of Electrical Engineering, Universitas Indonesia, from 1999 to 2004; where he was also the Head of the Telecommunication Laboratory, from 1997 to 2004. Since 2003, he has been the Director of the Antenna Propagation and Microwave Research Group (AMRG), Universitas Indonesia. He has published and presented more than 100 research articles in both national and international journals and symposiums. His research interests include antenna engineering, wave propagation, microwave circuits, communication systems, and telecommunication system regulations. He is a member of the IEEE Antenna and Propagation Society (AP-S) and the IEEE Microwave Theory and Technique Society (MTTS). He has been a member of the International Steering Committee (ISC) of the Asia Pacific Microwave Conference (APMC), since 2010, and the International Advisory Board of International Symposium on Antenna and Propagation (ISAP), since 2012. He was also the Head of the Telecommunication Laboratory, from 1997 to 2004. Since 2003, he has been the Director of the Antenna Propagation and Microwave Research Group (AMRG), Universitas Indonesia. He has published and presented more than 100 research articles in both national and international journals and symposiums. His research interests include antenna engineering, wave propagation, microwave circuits, communication systems, and telecommunication system regulations. He is a member of the IEEE Antenna and Propagation Society (AP-S) and the IEEE Microwave Theory and Technique Society (MTTS). He has been a member of the International Steering Committee (ISC) of the Asia Pacific Microwave Conference (APMC), since 2010, and the International Advisory Board of International Symposium on Antenna and Propagation (ISAP), since 2012. He was a recipient of the Indonesian Government Scholarship through MUCIA, from 1984 to 1987; the Hitachi Scholarship, from 1992 to 1996; the Young Researcher’s Award from Universitas Indonesia, in 1996; the Second Winner of Best Researcher Award in Science and Technology from Universitas Indonesia, in 2009; and the Second Winner of Best Teaching Award from Universitas Indonesia, in 2010. He was the Founder of the IEEE Joint Chapter MTT-S/AP-S Indonesia. He has served as the President for the IEEE Joint Chapter MTT-S/AP-S, from 2009 to 2010, and the IEEE Indonesia Section, from 2014 to 2015. He is the General Chairman of the Indonesia Microwave and Antenna Conference (IMMAC), Depok, in 2010; and the Indonesia Japan Joint Scientific Symposium (IJSS), Bali, in 2010. He was the General Co-Chairperson of the Indonesia Japan Joint Scientific Symposium (IJSS), Chiba, Japan, in 2012. In addition, he was also the General Chair of the 1st Indonesia Japan Workshop on Antennas and Wireless Technology (IJAWE), Depok, in 2017, and the 2019 IEEE International Conference on Antennas Measurements and Applications (CAMA), Bali, in October 2019.